AN-761
Application Note

VIDEO AMPLIFIER DESIGN: KNOW YOUR PICTURE TUBE REQUIREMENTS

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This note describes video amplifier design considerations for unitized gun and conventional picture tubes. Some unique design techniques are discussed taking advantage of Motorola's MC1323 chroma demodulator. Finally, design objectives of video amplifiers are discussed.

MOTOROLA Semiconductor Products Inc.

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INTRODUCTION

The advent of the unitized gun picture tube brought with it many claims of superior performance over the conventional delta gun configurations. Among these advantages were: improved gray scale tracking, better spot size and improved highlight resolution. These advancements were made possible because of the better grid-to-cathode cutoff ratios between guns. Although the combined gun structure (common G1 and G2 for all three guns) improved the grid-cathode cutoff ratio between guns, it now required the video amplifier output stage to compensate for any remaining differences in individual gun cutoffs. Thus, a simplification of the picture tube has resulted in a complication of the video stages driving it.

The video systems described in this note were designed to alleviate the design compromises normally associated with driving a unitized gun picture tube. These include interaction between driver and cutoff control, high power dissipation needed to obtain bandwidth, setup adjustment, dc stability and supply ripple rejection. Some of the designs can also be employed in conjunction with a conventional picture tube as very high-quality, wide-bandwidth, video systems.

BIASING REQUIREMENTS OF UNITIZED GUN TUBE

Cutoff characteristics for a typical unitized gun picture tube are presented in Figure 1. The two solid lines represent the spread in electron gun spot cutoff for all tubes. This variation is shown to be a ratio of 1.8-to-1 in G1-to-cathode voltage for any given G2 voltage. This ratio is greatly reduced for electron guns within a given tube to 1.2-to-1, as depicted by Tube A or Tube B. The shaded area marked Tube A shows the range of cutoff voltage for the three guns, assuming the one gun is a worst-case for maximum spot cutoff. Similarly, the shaded area marked Tube B shows the range of cutoff voltages assuming it contains a gun which is a worst-case for a minimum spot cutoff gun. The important point to be obtained from these curves is that the range of adjustment required by the video output stages to adjust spot cutoff will be the 1.2-to-1 ratio for any tube. For example, if it is desired to use a 150 Volt G1-to-cathode voltage, then a 25 Volt adjustment range is required. Notice the same range is required for Tube A as for Tube B, the only difference being the required G2 voltage. Thus, the large variation expected in cutoff ratios between tubes can be compensated for by adjusting the G2s.

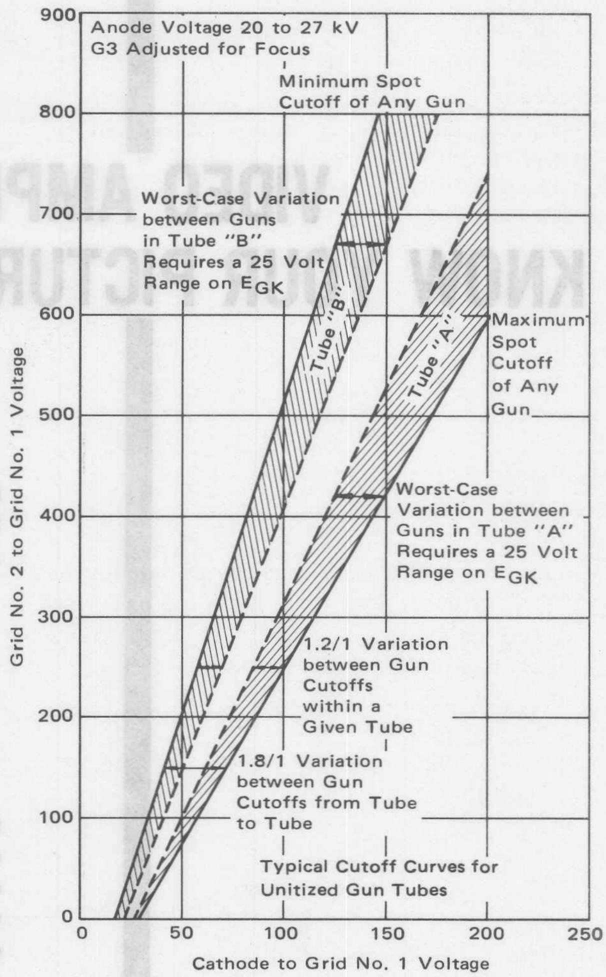


FIGURE 1

The cutoff ratio between guns in a given tube are compensated for by adjusting the cathode voltages. This method has the added advantage of using higher G2 voltages on tubes which do not have high cutoffs and thus obtain improved spot size. If, on the other hand, all tubes were biased at a fixed G2 of 420 Volts and the video amplifier used to compensate for cutoffs over the 1.8-to-1 ratio, the spot size on all tubes would be degraded to the worst-case condition. Also, in this case, the requirements on the video amplifier are more demanding.

Drive characteristic curves (Figure 2) show that with a cutoff of 150 Volts grid-to-cathode, a peak beam current of 6 mA per gun is possible. A gun with a 125 Volt E_{GK1} will still deliver 5 mA of peak current. Average current is limited between 1.0 and 3.0 mA, depending on the sys-

tem, by the Automatic Brightness Limiter (ABL) circuit to prevent damage to the picture tube. The high peak-to-average beam currents are necessary to reproduce peak whites and produce good highlight resolution.

To help design a video amplifier output stage consistent with the above requirements, it is useful to show the cathode and grid voltage on a diagram, as in Figure 3a. The G1 voltage should be selected as low as possible to keep the video output stage supply voltage low. This minimizes the power dissipation and voltage breakdown requirements of the video output devices. A lower limit on the G1 voltage is imposed by the saturation knee at high frequencies of the output stage. A common problem associated with setting the G1s too close to the saturation voltage of the output stage is color bleeding on luminance transients when accompanied by high color saturation levels. This problem manifests itself when the color sideband signals addition to the luminance in the output stage causes the luminance transients of one gun to be at a different level than the other gun, causing output stage saturation as shown in Figure 3b. Unitized gun tubes tend to exhibit this phenomenon, even without color drive, since the same effect is obtained when biasing the output stages at different black levels to compensate for the spot cutoffs of the guns. The effect of this can be avoided by: 1) picking the G1 voltage sufficiently above the saturation knee of the output devices, and 2) selection of a high

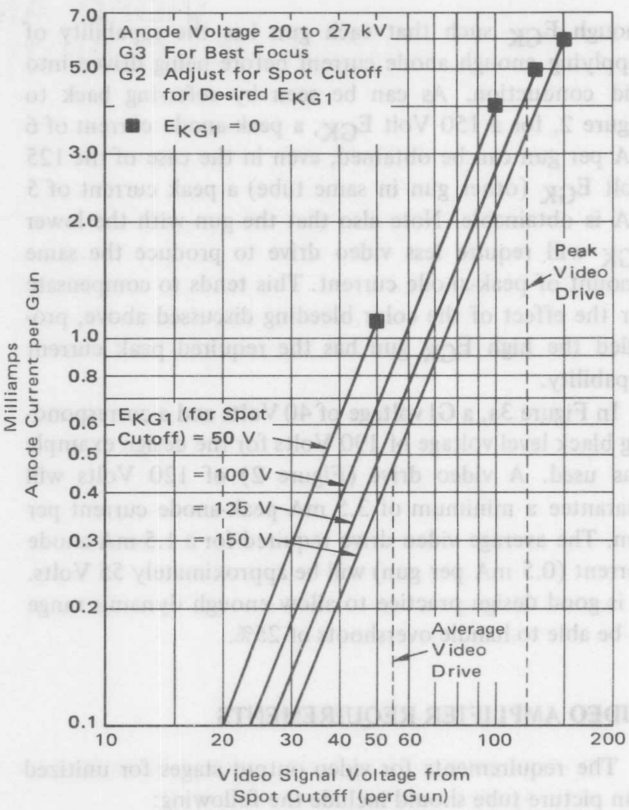


FIGURE 2 — Cathode Drive Characteristics

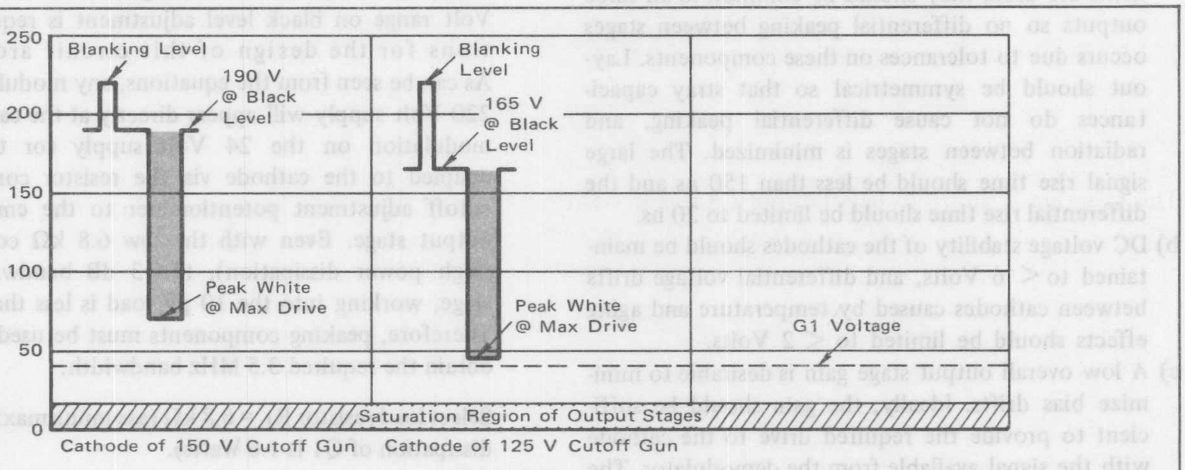


FIGURE 3a

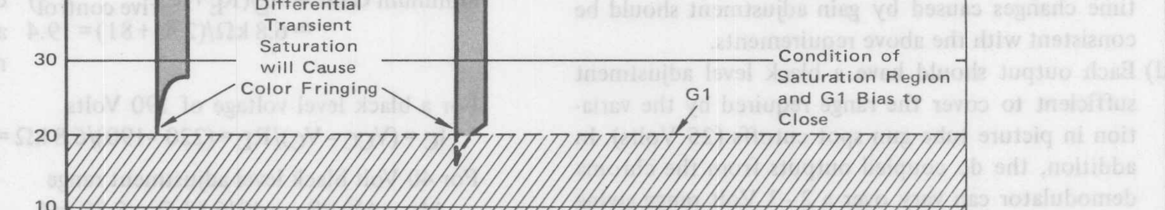


FIGURE 3b

FIGURES 3a and 3b — Video Output Stage Waveforms

enough E_{GK} such that each gun has the capability of supplying enough anode current before being driven into grid conduction. As can be seen by referring back to Figure 2, for a 150 Volt E_{GK}, a peak-anode current of 6 mA per gun can be obtained, even in the case of the 125 Volt E_{GK} (other gun in same tube) a peak current of 5 mA is obtainable. Note also that the gun with the lower E_{GK} will require less video drive to produce the same amount of peak-anode current. This tends to compensate for the effect of the color bleeding discussed above, provided the high E_{GK} gun has the required peak current capability.

In Figure 3a, a Gl voltage of 40 Volts and a corresponding black level voltage of 190 Volts for the design example was used. A video drive (Figure 2) of 120 Volts will guarantee a minimum of 3.5 mA peak-anode current per gun. The average video drive required for a 1.5 mA anode current (0.5 mA per gun) will be approximately 55 Volts. It is good design practice to allow enough dynamic range to be able to handle overshoots of 25%.

VIDEO AMPLIFIER REQUIREMENTS

The requirements for video output stages for unitized gun picture tube should include the following:

a) Bandwidth should be at least 3.5 MHz, either with or without peaking components. If peaking components are used, they should be common to all three outputs so no differential peaking between stages occurs due to tolerances on these components. Layout should be symmetrical so that stray capacitances do not cause differential peaking, and radiation between stages is minimized. The large signal rise time should be less than 150 ns and the differential rise time should be limited to 20 ns.

b) DC voltage stability of the cathodes should be maintained to $<$ 6 Volts, and differential voltage drifts between cathodes caused by temperature and aging effects should be limited to $<$ 2 Volts.

c) A low overall output stage gain is desirable to minimize bias drifts. Ideally, the gain should be sufficient to provide the required drive to the cathode with the signal available from the demodulator. The gain should be adjustable over a 30% range without affecting the black level voltage. Bandwidth and rise time changes caused by gain adjustment should be consistent with the above requirements.

d) Each output should have a black level adjustment sufficient to cover the range required by the variation in picture tube gun spot cutoffs (25 Volts). In addition, the dc coupled outputs from the chroma demodulator can vary over a 2–3 Volt range giving a 72 Volt variation in cathode voltage. If a demodulator device with a luminance/brightness input is used (MC1323 or MC1324), this cathode voltage variation can be reduced to 15 Volts (0.6 x 24). Thus, the black level adjustment control must accommodate at least a 40 Volt range and should not have any significant effects on gain or output stage bandwidth.

e) The transfer characteristics should be linear and independent of operating point and cutoff adjustment within the usable range (from black level to 0 Volts E_{GK}).

f) In order to maintain signal integrity, the rejection to power supply ripple should be maintained at $>$ 40 dB down from output signal. To obtain this, it is sufficient to either: 1) filter and regulate the power supply, or 2) design video output stages which have the required rejection to power supply modulation, or 3) a combination of both, to obtain the desired objective. Also note that the sensitivity of the output stages to power supply voltage should not cause more variation than the limits set for dc stability.

BRUTE FORCE APPROACH TO OBTAIN DESIGN OBJECTIVES

To begin the discussion, it is convenient to analyze a circuit for driving a unitized gun picture tube as illustrated in Figure 4. The design chosen requires operation with a 220 Volt B+ and a voltage gain of 24 to supply a 120 Volt drive with 5 volts of signal from the demodulator. As noted earlier, black level voltage of 190 Volts with a 40 Volt range on black level adjustment is required. Equations for the design of this circuit are included. As can be seen from the equations, any modulation on the 220 Volt supply will appear directly at the cathode. Also, modulation on the 24 Volt supply (or tolerance) is coupled to the cathode via the resistor connecting the cutoff adjustment potentiometer to the emitter of the output stage. Even with the low 6.8 kΩ collector load (high power dissipation), the 3 dB bandwidth of this stage, working into the 10 pF load is less than 2.5 MHz. Therefore, peaking components must be used in order to obtain the required 3.5 MHz bandwidth.

Select load resistor $R_L = 6.8\ k\Omega$ (assuming maximum power dissipation of Q1 is 1.6 Watts).

Maximum Gain $\simeq R_L/R_E \cong 6.8\ k\Omega/270 \simeq 25.2^*$ $\quad I_B << I_E$

Minimum Gain $\simeq R_L/(R_E + R_{drive\ control}) \simeq 6.8\ k\Omega/(270+81) = 19.4$

Select R_E and R_{drive} from desired max and min gain requirements.

For a black level voltage of 190 Volts

$I_C = (V_{CC} - V_C)/R_L = (220 - 190)/6.8\ k\Omega = 4.41$ mA

For 40 Volt black level adjustment range

$\Delta I_C = \Delta V_C/R_L = 40/6.8\ k\Omega = 5.88$ mA

assuming $\Delta I_C \simeq \Delta I_E$ and ΔI_E is to be obtained via cutoff control from 24 Volts.

$R_{EI} = 24/\Delta I_C \simeq 24/\Delta I_E \cong 24/5.88 \simeq 4\ k\Omega$

Quiesent dc voltage at demodulator output adjusted to 15

* Closest gain resulting from standard component values.

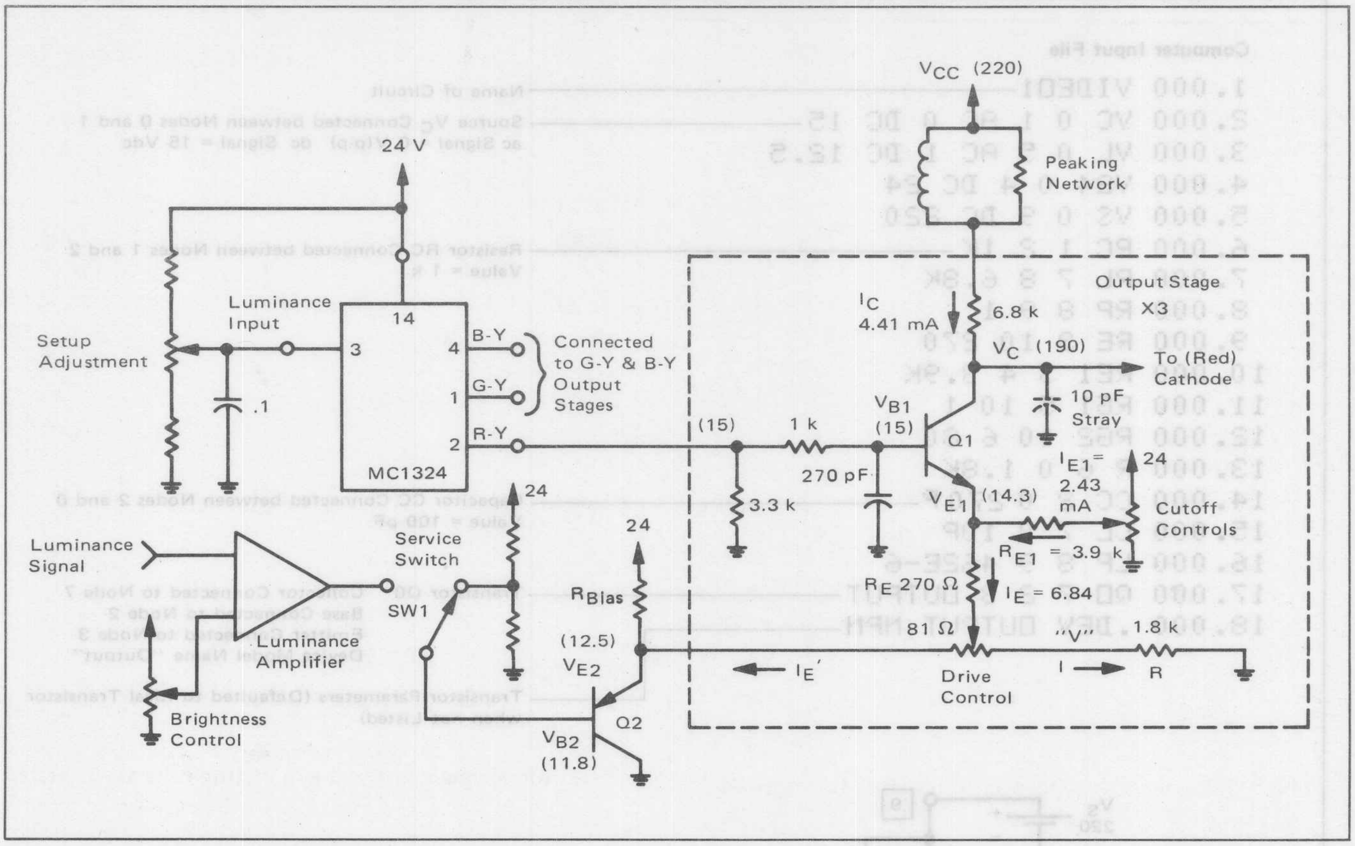


FIGURE 4

Volts on R-Y output via set-up adjustment.

$\therefore V_{B1} = 15$ Vdc (I_{B1} small) and $V_{E1} = 14.3$ Vdc
For Q1 on and $V_{BE} = .7$ Vdc

$I_{E1} = (24 - V_E)/R_{E1} = 2.43$ mA

(Note: if wiper is grounded $I_{E1} = (0 - V_E)/R_{E1} = -3.58$.
$\therefore \Delta I_{E1} = 2.43 - (-3.58) \simeq 5.8$ mA as predicted).

$I_E \simeq I_{E1} + I_C \simeq 4.41 + 2.43 \simeq 6.84$ mA assuming
$I_B << I_E$

Maximum Gain (drive control arm connected to emitter of video driver)

$V_{E1} - V_{E2} = I_E R_E = (6.84 \text{ mA})(270) = 1.85$
$V_{E2} = V_{E1} - I_E R_E = 14.3 - 1.85 = 12.5$
$V_{B2} = V_{E2} - V_{BE2} = 12.5 - 0.7 = 11.8$
$\therefore$ Bias network on base of video driver (V_{B2}) should be adjusted for 11.8 Vdc.

Minimum Gain (drive control arm connected to top of bridge resistor, R)

To prevent dc shift at cathode with drive control, maintain I_E @ 6.84 mA. Select R such that 6.84 mA will result in 12.5 Volts at "V".

$R = 12.5/6.84 \text{ mA} \simeq 1.8 \text{ k}\Omega$
$I_E' = (V - V_{E2})/R_{\text{drive control}} = (12.5 - 12.5)/81 = 0$
& $I_E = I + I_E' = 6.84 + 0 = 6.84$ mA

It should be emphasized that a demodulator with ability to set dc output must be used. A very serious drawback of this circuit is the inability to handle the variations in dc output voltage from one demodulator to the next. Typically, absolute output levels on demodulators vary 3 Volts from unit-to-unit, which would result in a 72 Volt tolerance at the cathode (3 Volts x gain of output stage). To correct this problem, a demodulator with a luminance input is employed and a potentiometer is used to adjust the quiescent output voltage to the required 15 Volts. An additional output variation between any two outputs to 0.6 Vdc can exist on a given demodulator. This variation, as previously discussed, can be compensated for at the cathode by the cutoff control, and at the expense of range on those controls. The 0.6 Volt offset also has the adverse effect of unbalancing the drive control bridge, thereby producing interaction between drive and cutoff controls. Motorola now has an MC1323 triple chroma demodulator with individually adjustable dc outputs, thus eliminating the need to compensate for the 0.6 Volt delta between outputs with the output stage. The range on control of the MC1323 dc outputs is also sufficient to compensate for picture tube cutoff variations $\frac{25 \text{ Volts}}{\text{gain of output stage}} \simeq 1$ Volt, thus eliminating the need for cutoff controls on the output stages. Refer to Application Note AN-763 for complete details on MC1323.

A quick computer analysis (using the Motorola Spice Circuit analysis program) of this basic circuit will aid in determining how well it meets design objectives. The model used and input documentation for this analysis is

Computer Input File

```
 1.000 VIDEO1
 2.000 VC 0 1 AC 0 DC 15
 3.000 VL 0 5 AC 1 DC 12.5
 4.000 V24 0 4 DC 24
 5.000 VS 0 9 DC 220
 6.000 RC 1 2 1K
 7.000 RL 7 8 6.8K
 8.000 RP 8 9 1
 9.000 RE 3 10 270
10.000 RE1 3 4 3.9K
11.000 RG1 5 10 1
12.000 RG2 10 6 80
13.000 R 6 0 1.8K
14.000 CC 2 0 270P
15.000 CL 7 0 10P
16.000 LP 8 9 462E-6
17.000 Q0 7 2 3 OUTPUT
18.000 .DEV OUTPUT NPN
```

Name of Circuit

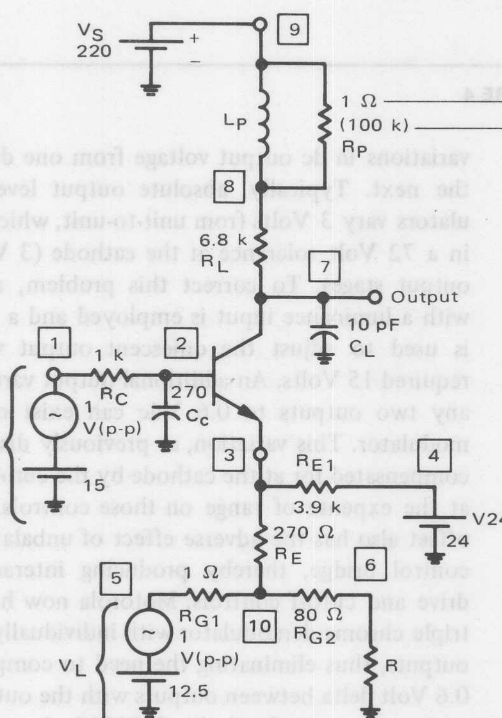
Source V_C Connected between Nodes 0 and 1
ac Signal = 0 V(p-p) dc Signal = 15 Vdc

Resistor RC Connected between Nodes 1 and 2
Value = 1 k

Capacitor CC Connected between Nodes 2 and 0
Value = 100 pF

Transistor Q0 Collector Connected to Node 7
Base Connected to Node 2
Emitter Connected to Node 3
Device Model Name "Output"

Transistor Parameters (Defaulted to Ideal Transistor when not Listed)



Value used to Effectively Short Peaking Coil

Value used when "Sweeping" to Show Effects of Peaking

Varied from 24 Vdc to 0 Vdc for Cutoff Control Range

V_C Chroma Sideband Source
V_L Luminance Signal Source
R_{G1} & R_{G2} Gain Control
Max Gain R_{G1} = 1, R_{G2} = 80
Min Gain R_{G1} = 80, R_{G2} = 1

FIGURE 5a – Circuit Model and Input File Description

shown in Figure 5a. The computed node voltages, supply currents, transistor operating point, and element sensitivity chart are included in Figure 5b. The sensitivities are expressed in volts per unit (change in output voltage for a one unit change in element value) and volts per percent (change in output voltage for a one percent change in element value). This table will help in selecting tolerances on components as well as setting requirements for our 220 Volt power supply. For example, if our minimum (low contrast) output signal is 60 V(p-p), and we desire a power supply ripple rejection of > 40 dB in our output signal, the output ripple must be ⩽ 0.6 V(p-p). The V_s supply sensitivity results in a 1 Volt change in output voltage for a 1 Volt change in supply, thus the V_s supply

```
NODE  VOLTAGE        NODE  VOLTAGE        NODE  VOLTAGE        NODE  VOLTAGE
(  1)   15.0000     (  2)   14.9598      (  3)   14.2686     (  4)   24.0000
(  5)   12.5000     (  6)   14.9680      (  7)  192.6954     (  8)  220.0000
(  9)  220.0000     ( 10)   12.4999

VOLTAGE SOURCE CURRENTS

NAME        CURRENT

VC         4.015E-05 AMPS
VL         9.811E-05 AMPS
V24        2.495E-03 AMPS
VS         4.015E-03 AMPS

TOTAL POWER DISSIPATION    9.45E-01  WATTS


TRANSISTOR OPERATING POINTS

NAME      MODEL       IB          IC          VBE      VBC      VCE      BETADC

QO        OUTPUT    4.96E-05   4.96E-03      .691   177.736  178.427    100.0

SMALL SIGNAL CHARACTERISTICS

V7       /  VL           INPUT IMPEDANCE          OUTPUT IMPEDANCE
                           AT   VL                    AT   V7
   2.344E 01               2.485E 02                 6.800E 03

    NAME            VALUE            SENSITIVITY          SENSITIVITY
                                   (VOLTS/ UNIT)       (VOLTS/PERCENT)

    RC            1.000E 03         1.007E-03             1.007E-02
    RL            6.800E 03        -4.015E-03            -2.730E-01
    RP            1.000E 00          .000E 00              .000E 00
    RE            2.700E 02         1.536E-01             4.148E-01
    RE1           3.900E 03        -4.066E-03            -1.586E-01
    RG1           1.000E 00        -2.299E-03            -2.299E-05
    RG2           8.000E 01         8.289E-05             6.631E-05
    R             1.800E 03         8.289E-05             1.492E-03
    VC            1.500E 01        -2.508E 01            -3.762E 00
    VL            1.250E 01         2.344E 01             2.930E 00
    V24           2.400E 01         1.629E 00             3.911E-01
    VS            2.200E 02         1.000E 00             2.200E 00
```

FIGURE 5b – Black Level dc Characteristics

ripple must be < 0.6 Volts. The sensitivity of the V24 supply is 1.63 Volts per Volt, requiring a ripple of < 0.6/1.63 or 0.37 V(p-p) to maintain signal integrity. The absolute value of the supplies must be regulated against line and load, due to the high sensitivity.

The computed dc transfer characteristics are shown in Figure 5c for minimum and maximum cutoff settings. Notice that in each case the solid and dotted lines are parallel, illustrating that the cutoff adjustment does not change the gain. Also, the solid and dotted lines cross at black level, indicating the gain adjustment does not affect the cutoff. The range of cutoff adjustment from the curves is 192 Volts minus 153 Volts, or 39 Volts. The gain control range varies from a maximum of 23.4 (slope of solid lines) to a minimum of 17.9 (slope of dotted lines). It is always reassuring to compare these computed results with those of the initial design.

The computed response curves for both the luminance and chroma sideband signals are shown in Figure 5d. The discrepancy in dc gain is caused by the voltage divider formed by R_{E1} and R_E when driving from the luminance input, but which does not attenuate the chroma sideband signal. This effect is minimal as long as $R_{E1} >> R_E$. In both cases, the effective emitter resistance is the parallel

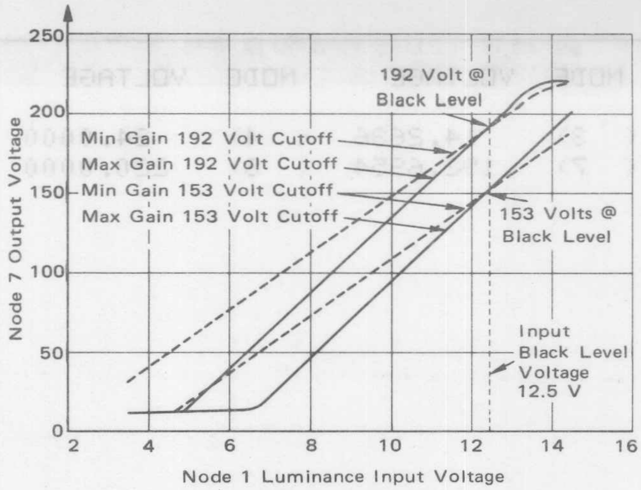


FIGURE 5c

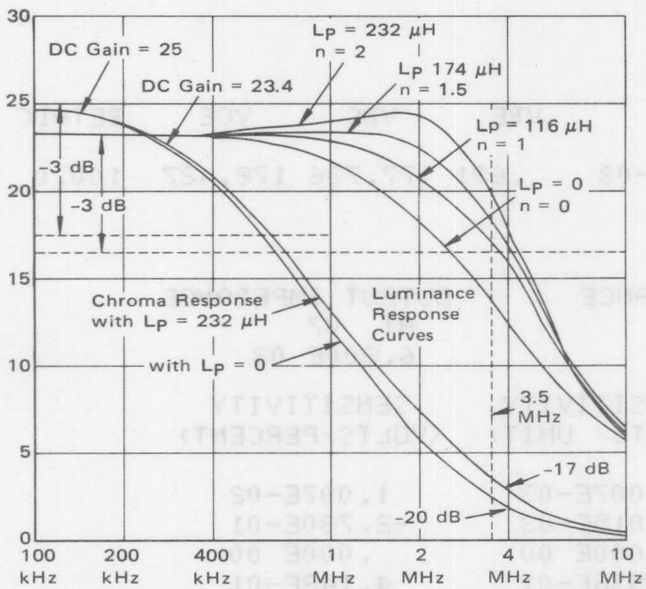


FIGURE 5d

combination of R_{E1} and R_E. The video response is shown for various values of Lp where $Lp = n\ R_L{}^2\ C_L/4$ and n = 0, 1, 1.5, 2. As can be seen from the curves, maximum bandwidth with no response peaking occurs for n = 1. (The poles of the transfer function are real and equal for n = 1). In practice, for values of $n > 2$, it is common to use some damping resistance across Lp to prevent ringing. The effect of the peaking inductance on the chroma response is shown for the extreme values of n.

In summary, this type of output stage will satisfy the requirements for driving a unitized gun picture tube but it does put some rather stringent requirements on the power supply. It also requires a high power output device because of the low value of R_L.

NOVEL SOLUTIONS TO DESIGN OBJECTIVES

An operational amplifier with feedback has many properties which lend themselves very nicely to video output stages. These include: very low output impedance, accurately defined gain, high power supply rejection ratio, summing input capability and performance characteristics generally independent of the amplifier parameters. Unfortunately, operational amplifiers are not suited for 250 Volt operation with 150 Volt swing capability. The objective here is to use components normally used in video output stages, arrange them in a high gain configuration (pseudo operational amplifier) and apply operational amplifier theory to take advantage of the above mentioned characteristics.

In it's simplest form, the pseudo operational amplifier of the improved video system is shown in Figure 6a. The voltage gain of this section is very high and can be shown to be $A_V = \frac{R_L}{R_e + r_e}$ where r_e = 26 mV/I_E at room temperatures. With the values shown $A_V = 435$.

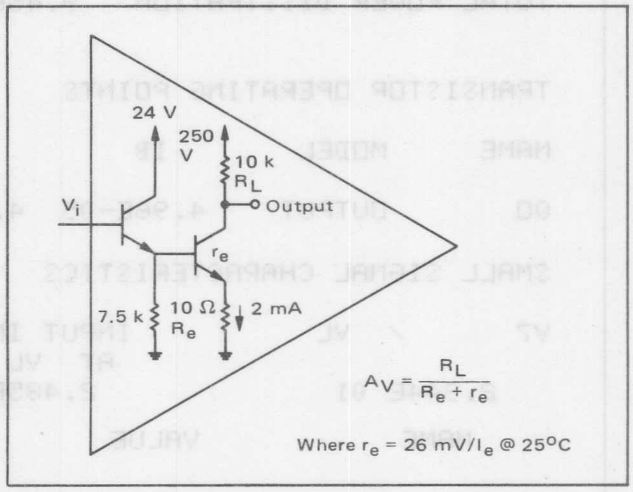


FIGURE 6a

If the above figure is redrawn in operational amplifier configuration, it can be utilized as a summing amplifier. By selecting the ratio of R_F to R_{in} the gain to each input can be individually selected.

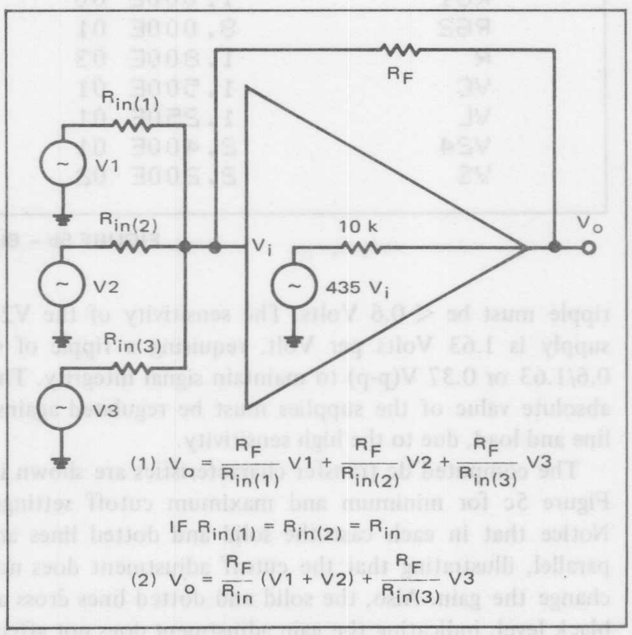


FIGURE 6b

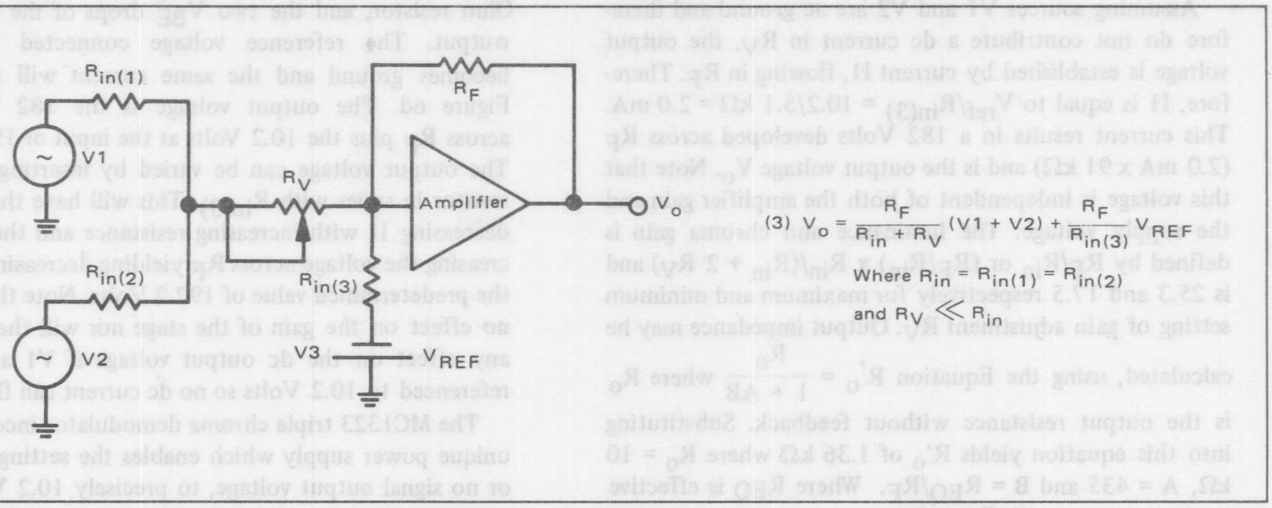


FIGURE 6c

Replacing V1 with the luminance signal Y1 and V2 with the demodulated R-Y chroma sideband signals will result in the required matrix for red, required to drive the red cathode.

Two other similar stages are used to obtain the blue and green outputs respectively. The third input V3 is used to establish the black level voltage at V_O which is connected to the picture tube cathodes. As can be seen from Equation 2 in Figure 6b, the dc voltage at V_O can be changed by varying $R_{in(3)}$ with a dc reference voltage substituted for V3. Also the gain to V1 and V2 can be ganged together by the use of a common variable resistor in their signal path. Figure 6c illustrates these modifications.

Substituting some typical values in Figure 6c, gain, dc output voltage and output impedance can be calculated. Refer to Figure 6d for values.

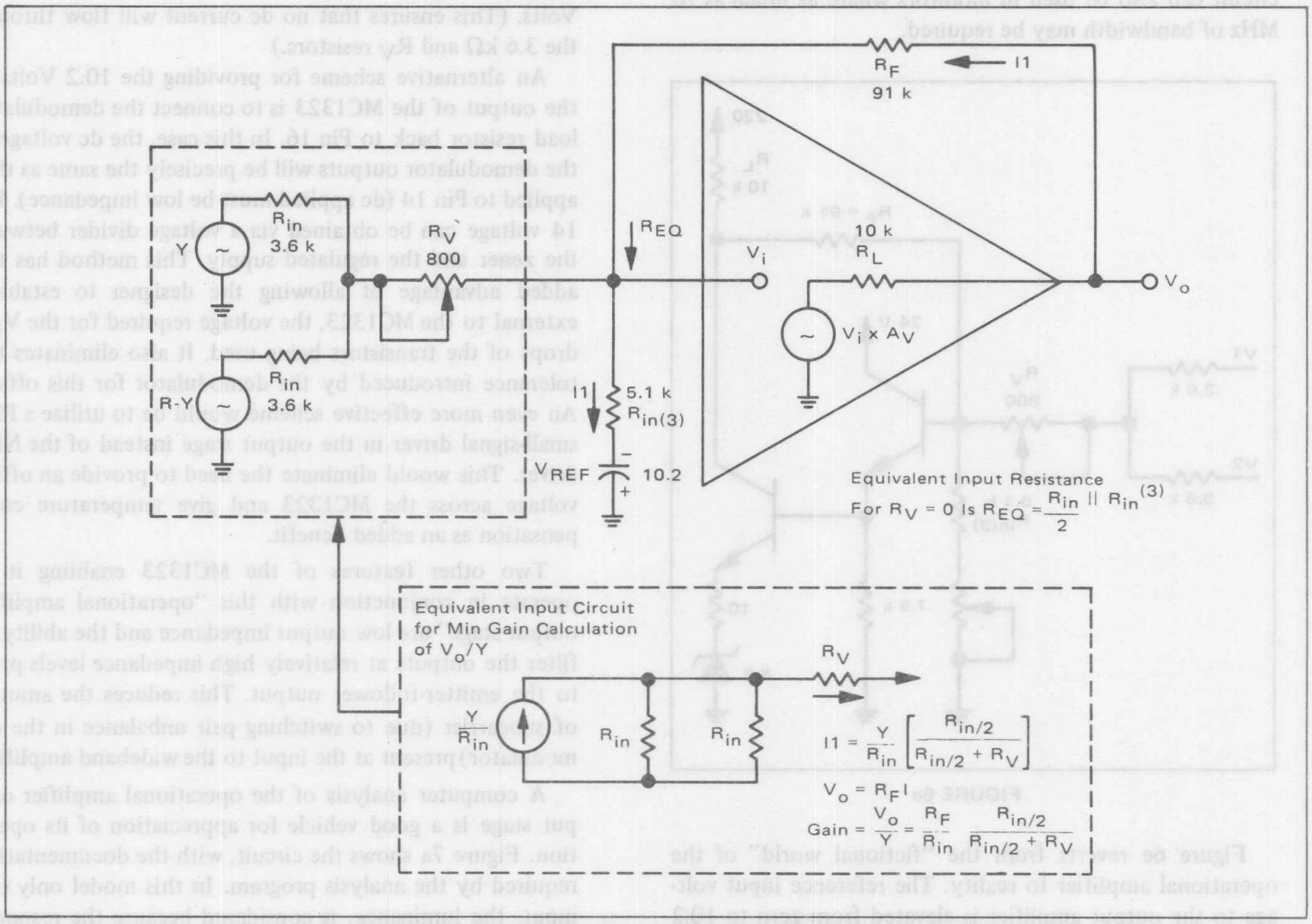


FIGURE 6d

Assuming sources V1 and V2 are ac ground and therefore do not contribute a dc current in R_V, the output voltage is established by current I1, flowing in R_F. Therefore, I1 is equal to $V_{ref}/R_{in(3)} = 10.2/5.1\ k\Omega = 2.0\ mA$. This current results in a 182 Volts developed across R_F (2.0 mA x 91 kΩ) and is the output voltage V_o. Note that this voltage is independent of both the amplifier gain and the supply voltage. The luminance and chroma gain is defined by R_F/R_{in} or $(R_F/R_{in}) \times R_{in}/(R_{in} + 2\ R_V)$ and is 25.3 and 17.5 respectively for maximum and minimum setting of gain adjustment R_V. Output impedance may be calculated, using the Equation $R'_o = \frac{R_o}{1 + AB}$ where R_o is the output resistance without feedback. Substituting into this equation yields R'_o of 1.36 kΩ where R_o = 10 kΩ, A = 435 and B = R_{EQ}/R_F. Where R_{EQ} is effective input resistance with R_V = 0.

The effect of this low output resistance working into a 10 pF capacitive load (output capacity of transistor plus picture tube cathode capacity plus stray of wire from output stage to picture tube) gives a bandwidth well in excess of the 4 MHz required. This gives the advantage of not needing to peak the low level video amplifiers to compensate for frequency roll-off in the video output stage. It is also easier to produce high-frequency peaking which will give narrower overshoots and a crisper picture. This circuit can also be used in monitors where as much as 10 MHz of bandwidth may be required.

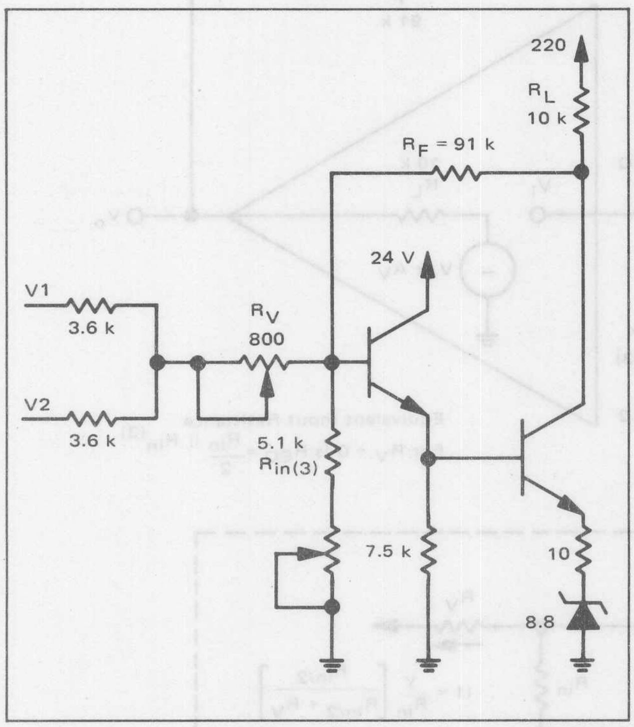


FIGURE 6e

Figure 6e reverts from the "fictional world" of the operational amplifier to reality. The reference input voltage to the output amplifier is elevated from zero to 10.2 Volts via the zener diode the voltage drop across the 10 Ohm resistor, and the two V_{BE} drops of the darlington output. The reference voltage connected to $R_{in(3)}$ becomes ground and the same current will flow as in Figure 6d. The output voltage is the 182 Volt drop across R_F plus the 10.2 Volts at the input or 192.2 Volts. The output voltage can be varied by inserting a variable resistor in series with $R_{in(3)}$. This will have the effect of decreasing I1 with increasing resistance and therefore decreasing the voltage across R_F yielding decreasing V_o from the predetermined value of 192.2 Volts. Note that this has no effect on the gain of the stage nor will the gain have any effect on the dc output voltage if V1 and V2 are referenced to 10.2 Volts so no dc current can flow in R_V.

The MC1323 triple chroma demodulator incorporates a unique power supply which enables the setting of V1 dc, or no signal output voltage, to precisely 10.2 Volts without the use of a "setup" potentiometer. Basically, as shown in Figure 6f, the dc output of the MC1323 is set 1.6 Volts above the voltage applied to Terminal 14 by referencing the three demodulator loads back to Pin 15. (See MC1323 data sheet for complete description of power supply operation.) Using the 8.8 Volt zener diode as a reference to the MC1323 power supply (Pin 14), the demodulated R-Y, B-Y and G-Y signals will be referenced 1.6 Volts above the zener. In practice, the two V_{BE} voltage drops plus the drop across the 10-Ohm resistor will be approximately 1.6 Volts so the bridge voltage will be 10.4 Volts. (This ensures that no dc current will flow through the 3.6 kΩ and R_V resistors.)

An alternative scheme for providing the 10.2 Volts at the output of the MC1323 is to connect the demodulator load resistor back to Pin 16. In this case, the dc voltage at the demodulator outputs will be precisely the same as that applied to Pin 14 (dc applied must be low impedance). Pin 14 voltage can be obtained via a voltage divider between the zener and the regulated supply. This method has the added advantage of allowing the designer to establish external to the MC1323, the voltage required for the V_{BE} drops of the transistors being used. It also eliminates the tolerance introduced by the demodulator for this offset. An even more effective scheme would be to utilize a PNP small-signal driver in the output stage instead of the NPN driver. This would eliminate the need to provide an offset voltage across the MC1323 and give temperature compensation as an added benefit.

Two other features of the MC1323 enabling it to operate in conjunction with this "operational amplifier output stage" are low output impedance and the ability to filter the outputs at relatively high impedance levels prior to the emitter-follower output. This reduces the amount of subcarrier (due to switching pair unbalance in the demodulator) present at the input to the wideband amplifier.

A computer analysis of the operational amplifier output stage is a good vehicle for appreciation of its operation. Figure 7a shows the circuit, with the documentation required by the analysis program. In this model only one input, the luminance, is considered because the response to the chroma sidebands will be identical due to the cir-

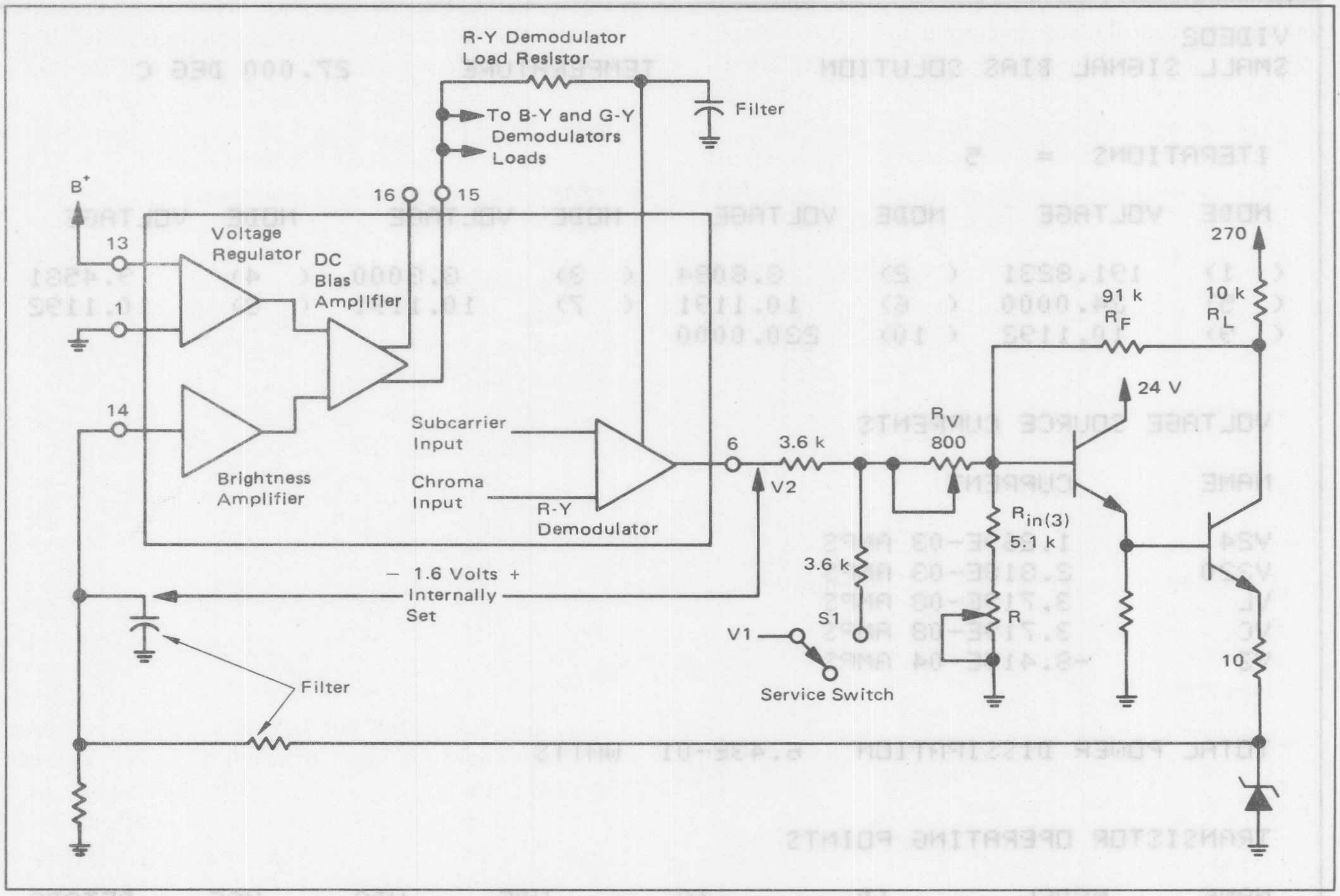


FIGURE 6f

```
VIDEO2
V24 0 5 24
V220 0 10 DC 220
VL 0 9 AC 1 DC 10.1192
VC 0 8 DC 10.1192
VZ 0 3 DC 8.8
RIN1 9 7 3.6K
RIN2 8 7 3.6K
RV 7 6 1
RIN3 6 0 5.1K
RF 1 6 91K
RE1 4 0 7.5K
RE 2 3 10
RL 10 1 10K
Q1 1 4 2 PWR
Q2 5 6 4 SS
.DEV PWR NPN BF=40 CJE=30P CJC=10P
.DEV SS NPN
.OUTPUT V1 V(1,0) PRI MA PH DC
```

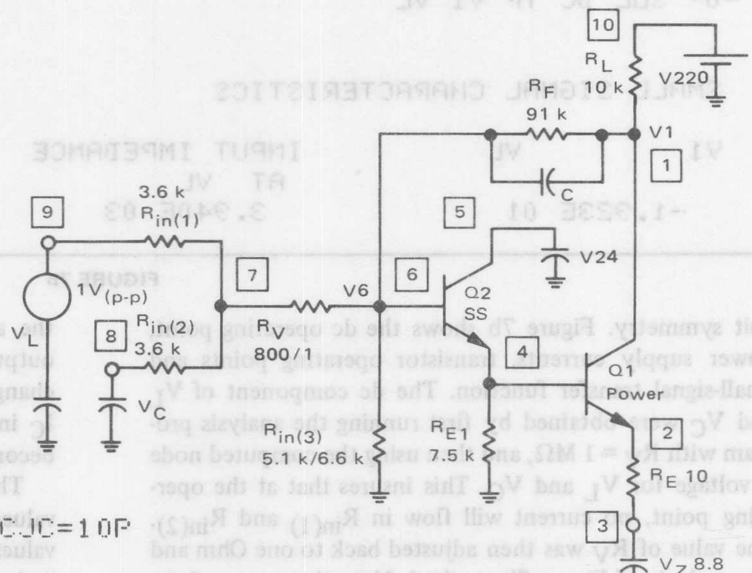


FIGURE 7a

```
VIDEO2
SMALL SIGNAL BIAS SOLUTION              TEMPERATURE      27.000 DEG C

 ITERATIONS  =    5

 NODE   VOLTAGE       NODE   VOLTAGE       NODE   VOLTAGE       NODE   VOLTAGE

(  1)    191.8231   (  2)      8.8084   (  3)      8.8000 (  4)       9.4581
(  5)     24.0000   (  6)     10.1191   (  7)     10.1191 (  8)      10.1192
(  9)     10.1192   ( 10)    220.0000

 VOLTAGE SOURCE CURRENTS

 NAME          CURRENT

 V24           1.269E-03 AMPS
 V220          2.818E-03 AMPS
 VL            3.719E-08 AMPS
 VC            3.719E-08 AMPS
 VZ           -8.415E-04 AMPS

 TOTAL POWER DISSIPATION    6.43E-01  WATTS

 TRANSISTOR OPERATING POINTS

 NAME      MODEL          IB         IC         VBE      VBC       VCE      BETADC

 Q1        PWR         2.19E-05  8.77E-04     .650  182.365  183.015     40.0
 Q2        SS          1.33E-05  1.33E-03     .661  -13.881   14.542    100.0

-0- SOL DC TF V1 VL

SMALL SIGNAL CHARACTERISTICS

V1          VL              INPUT IMPEDANCE          OUTPUT IMPEDANCE
                             AT  VL                    AT  V1
  -1.923E 01                   3.940E 03                 2.151E 03
```

FIGURE 7b

cuit symmetry. Figure 7b shows the dc operating point, power supply currents, transistor operating points and small-signal transfer function. The dc component of V_L and V_C were obtained by first running the analysis program with $R_V = 1\ M\Omega$, and then using the computed node 6 voltage for V_L and V_C. This insures that at the operating point, no current will flow in $R_{in(1)}$ and $R_{in(2)}$. The value of R_V was then adjusted back to one Ohm and the analysis of Figure 7b resulted. Note that some of the calculation presented earlier will yield slightly different results than those of the computer analysis. These differences result from the assumption that the output stage was a perfect operational amplifier. The computer analysis does show, however, that only small errors resulted from the approximations and that this is a very useful video output stage. (The major cause of the discrepancies is the change in input voltage at Pin 6 due to V_{BE} changes with I_C in the output transistor. Note that the effect of this becomes small as the output stage current increases.)

The sensitivity of the output dc voltage to element values is tabulated in Figure 7c. Comparison of these values with those computed for the previous circuit show a significant decrease in power supply sensitivity. Stable resistors with similar temperature coefficient should be used to minimize output drift. The absolute value is not critical as $R_{in(3)}$ will have a variable resistor in series with it for black level adjustment. Sensitivity to the zener is somewhat compensated for by the black level adjustment

```
VIDEO2
DC SENSITIVITIES                                TEMPERATURE        27.000 DEG C

SENSITIVITIES OF V1
```

PART NAME	PART VALUE	PART SENSITIVITY (VOLTS/ UNIT)	NORMALIZED SENSITIVITY (VOLTS/PERCENT)
RIN1	3.600E 03	7.152E-07	2.575E-05
RIN2	3.600E 03	7.152E-07	2.575E-05
RV	1.000E 00	2.861E-06	2.861E-08
RIN3	5.100E 03	-2.695E-02	-1.374E 00
RF	9.100E 04	1.567E-03	1.426E 00
RE1	7.500E 03	-2.868E-04	-2.151E-02
RE	1.000E 01	4.465E-02	4.465E-03
RL	1.000E 04	-6.061E-04	-6.061E-02
V24	2.400E 01	-9.673E-08	-2.322E-08
V220	2.200E 02	2.151E-01	4.733E-01
VL	1.012E 01	-1.923E 01	-1.946E 00
VC	1.012E 01	-1.923E 01	-1.946E 00
VZ	8.800E 00	5.306E 01	4.669E 00

FIGURE 7c

potentiometer and also by referencing the dc component of V_C to the zener voltage via the demodulator as discussed previously. The high sensitivity to the zener might be a concern and will be given more rigorous treatment later.

The analysis was repeated for $R_{in(3)}$ equal to 6.6 kΩ and the node voltages, transistor operating points, etc., are tabulated in the computer printout in Figure 7d. Comparison of the data for the two different values of $R_{in(3)}$ shows a change in gain $V1/V_L$ at the operating point. This is a result of the low current in the output transistor in the case of $R_{in(3)}$ = 5.1 kΩ. As the signal V_L is applied and current increases in the output stage, the gain will increase to the same value as that shown for $R_{in(3)}$ = 6.6 kΩ. This effect is indicated in Figure 7e, which represents the transfer characteristic of the amplifier for the two values of $R_{in(3)}$. The data used to plot these curves along with the small-signal gain and impedance levels are presented in Figure 7f. The lower gain at low-current levels, again represent the departure of the configuration used from a true operational amplifier. The voltage gain is decreasing and the gain is no longer a function of the ratio of the resistors. This effect and the decrease in sensitivity to supply voltage is explained as follows: $G_M = I_E/26$ mV at room temperature increases with increasing I_E. Both lower output voltage and/or higher supply voltage will cause an increase in I_E. The increase in I_E causes the voltage sensitivity at the base of Q1 to decrease ($gm = \Delta I_E/\Delta V_{BE}$ or $\Delta V_{BE} = \Delta I_E/gm$) therefore, reflecting a decrease in sensitivity at the base of Q2 and finally at the output after being multiplied by the ratio of the feedback resistors. The supply voltage can be raised above 220 Volts to improve gain linearity and decrease supply sensitivity. $R_{in(1)}$ and $R_{in(2)}$ have current flowing in them due to the increased voltage at the base of Q2 causing their sensitivities to increase from the previous computer run in which $R_{in(3)}$ = 5.1 kΩ. Note, however, that the sensitivities are still quite small.

A computer analysis of the frequency response of this stage substantiates the claims previously made for it. Figure 7g shows the effects of gain and bias point variation on the frequency response. The capacitor C is included to show that the feedback resistor cannot be chosen arbitrarily high, in order to minimize current through the load resistor required for the feedback. The end-to-end capacity of a 1/2-Watt resistor is ≈ 0.4 pF worst-case. The 91 kΩ resistor used will cause a transmission zero to occur at ($f = 1/2\pi RC$) 4.4 MHz. Higher values of R_F will bring the zero into the video passband. Smaller values of C will yield response curves somewhere between the limits shown. The computer data is tabulated in Figure 7h.

It is important to note that the C_{cb} of the small-signal driver transistor is not in parallel with R_F but is reflected as a capacity to ground because of the common collector configuration. Appendix A indicates the small-signal current gain transfer characteristics for the model used for the output transistor. The purpose for including this material is twofold: a) to justify that the model used accurately predicts the response characteristics of a typical high-voltage transistor used for output stages in TV receivers and, b) to compare with the results of the response shown in Figure 7g.

A sweep response of this amplifier was performed in the laboratory and the results are shown in Figure 7i, with a 10 kHz and 1 MHz square wave response. Notice that

both the rise and fall times of the square wave are < 100 ns without peaking components used. Some of the values used in the actual circuit are slightly different from the preceding analysis in order to satisfy the requirements of a particular chassis. The variations will not cause any significant differences in results. As a comparison, Figure 7j is included to show the response of a "standard" output stage. As can be seen, the response even with peaking does not have sufficient bandwidth for a high-grade system and peaking in the low-level video amplifier is normally employed to compensate for the output stage roll-off.

```
VIDEO2
SMALL SIGNAL BIAS SOLUTION             TEMPERATURE        27.000 DEG C

 ITERATIONS  =    5

 NODE  VOLTAGE        NODE  VOLTAGE         NODE  VOLTAGE         NODE  VOLTAGE

(  1)  156.5956    (  2)      8.8485    (  3)      8.8000    (  4)      9.5434
(  5)   24.0000    (  6)     10.2065    (  7)     10.2064    (  8)     10.1192
(  9)   10.1192    ( 10)    220.0000

 VOLTAGE SOURCE CURRENTS

 NAME       CURRENT

 V24         1.377E-03 AMPS
 V220        6.340E-03 AMPS
 VL         -2.423E-05 AMPS
 VC         -2.423E-05 AMPS
 VZ         -4.850E-03 AMPS

 TOTAL POWER DISSIPATION    1.38E 00   WATTS

 TRANSISTOR OPERATING POINTS

 NAME    MODEL       IB         IC              VBE     VBC      VCE     BETADC

 Q1      PWR      1.18E-04   4.73E-03         .695  147.052  147.747    40.0
 Q2      SS       1.43E-05   1.43E-03         .663  -13.794   14.457   100.0

  SMALL SIGNAL CHARACTERISTICS

 V1      /    VL          INPUT IMPEDANCE         OUTPUT IMPEDANCE
                               AT  VL                   AT  V1
    -2.255E 01               3.753E 03               9.669E 02

DC SENSITIVITIES                          TEMPERATURE        27.000 DEG C

  SENSITIVITIES OF V1

      PART           PART               PART           NORMALIZED
      NAME           VALUE           SENSITIVITY       SENSITIVITY
                                    (VOLTS/ UNIT)    (VOLTS/PERCENT)

      RIN1        3.600E 03          -5.465E-04        -1.967E-02
      RIN2        3.600E 03          -5.465E-04        -1.967E-02
      RV          1.000E 00          -2.186E-03        -2.186E-05
      RIN3        6.600E 03          -1.903E-02        -1.256E 00
      RF          9.100E 04           1.453E-03         1.322E 00
      RE1         7.500E 03          -3.140E-04        -2.355E-02
      RE          1.000E 01           2.840E-01         2.840E-02
      RL          1.000E 04          -6.131E-04        -6.131E-02
      V24         2.400E 01          -1.134E-07        -2.721E-08
      V220        2.200E 02           9.669E-02         2.127E-01
      VL          1.012E 01          -2.255E 01        -2.282E 00
      VC          1.012E 01          -2.255E 01        -2.282E 00
      VZ          8.800E 00           5.856E 01         5.154E 00
```

FIGURE 7d

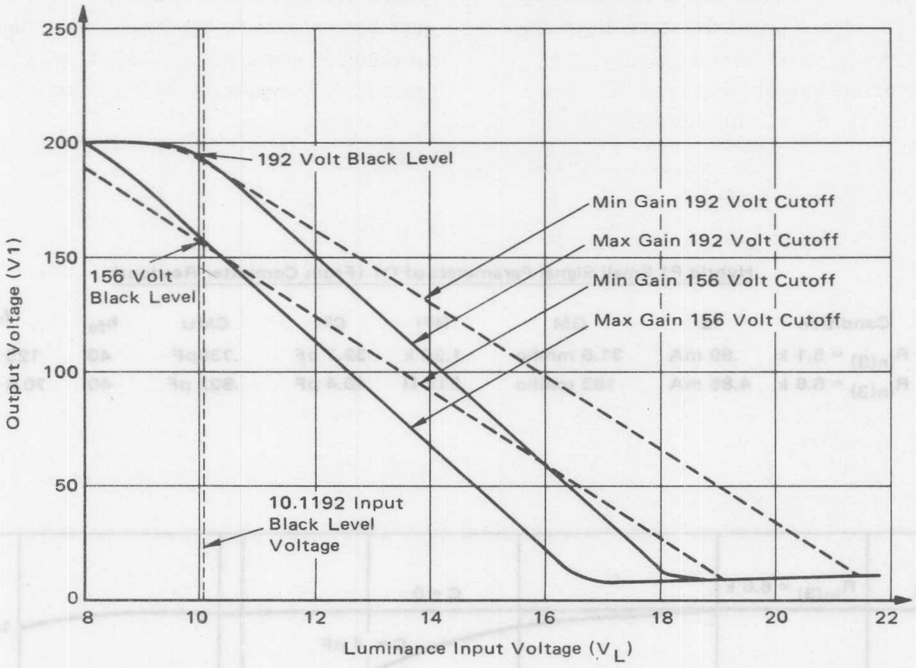


FIGURE 7e

VIDEO2

DC TRANSFER CURVES   TEMPERATURE 27.000 DEG C

Luminance Input VL	$R_{in(1)}$ = 5.1 k, R_V = 1 Ω V1	$R_{in(3)}$ = 5.1 k, R_V = 800 V1	$R_{in(3)}$ = 6.6 k, R_V = 800 V1	$R_{in(3)}$ = 6.6 k, R_V = 1Ω V1
8.00E 00	1.992E 02	1.992E 02	1.886E 02	1.988E 02
9.00E 00	1.992E 02	1.992E 02	1.732E 02	1.814E 02
1.00E 01	1.940E 02	1.935E 02	1.573E 02	1.593E 02
1.10E 01	1.734E 02	1.787E 02	1.412E 02	1.366E 02
1.20E 01	1.512E 02	1.631E 02	1.250E 02	1.137E 02
1.30E 01	1.285E 02	1.471E 02	1.088E 02	9.069E 01
1.40E 01	1.057E 02	1.311E 02	9.248E 01	6.760E 01
1.50E 01	8.277E 01	1.149E 02	7.617E 01	4.446E 01
1.60E 01	5.978E 01	9.877E 01	5.983E 01	2.128E 01
1.70E 01	3.674E 01	8.256E 01	4.348E 01	9.119E 00
1.80E 01	1.366E 01	6.632E 01	2.712E 01	9.248E 00
1.90E 01	9.153E 00	5.006E 01	1.074E 01	9.396E 00
2.00E 01	9.287E 00	3.378E 01	9.146E 00	9.551E 00
2.10E 01	9.434E 00	1.749E 01	9.251E 00	9.708E 00
2.20E 01	9.585E 00	9.109E 00	9.366E 00	9.866E 00
SMALL SIGNAL CHARACTERISTICS				
V1 / VL	-1.923E 01	-1.409E 01	-1.603E 01	-2.255E 01
OUTPUT IMPEDANCE AT V1	2.151E 03	1.757E 03	7.546E 02	9.669E 02
INPUT IMPEDANCE AT VL	3.940E 03	4.484E 03	4.356E 03	3.753E 03

FIGURE 7f

Hybrid P1 Small Signal Parameters of Q1 (From Computer Readout)

Condition	IE	GM	RPI	CPI	CMU	h_{fe}	f_T
$R_{in(3)}$ = 5.1 k	.89 mA	31.8 mmho	1.26 k	39.7 pF	.738pF	40	125 MHz
$R_{in(3)}$ = 6.6 k	4.85 mA	183 mmho	219 Ω	40.4 pF	.822 pF	40	70.6 MHz

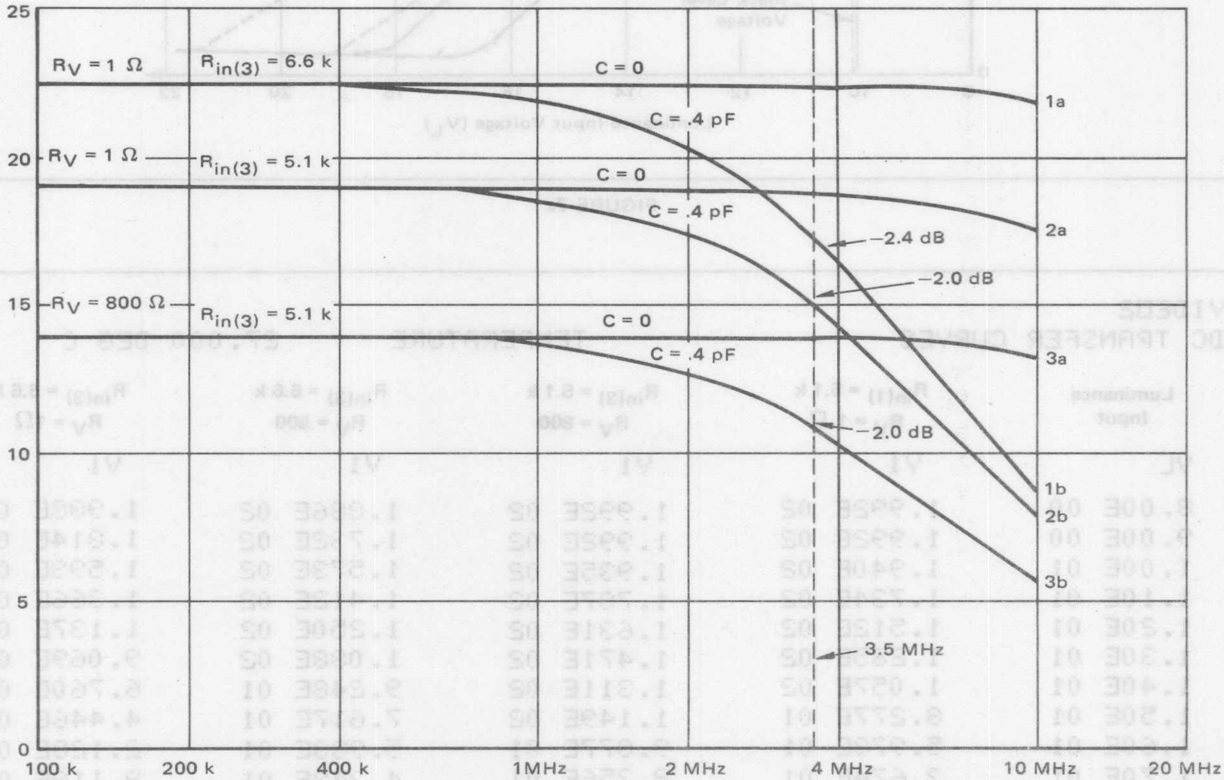
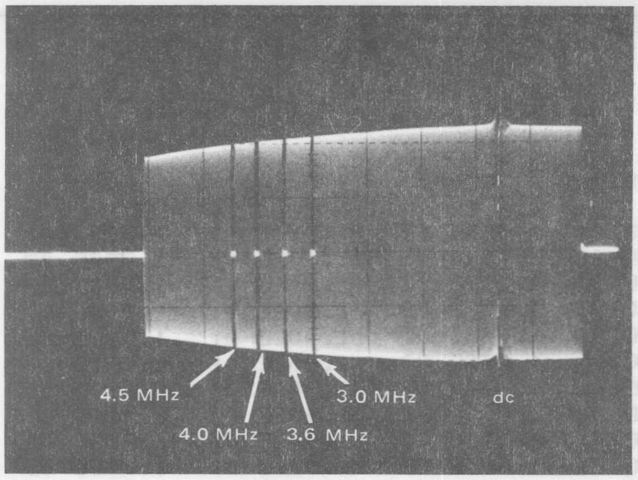


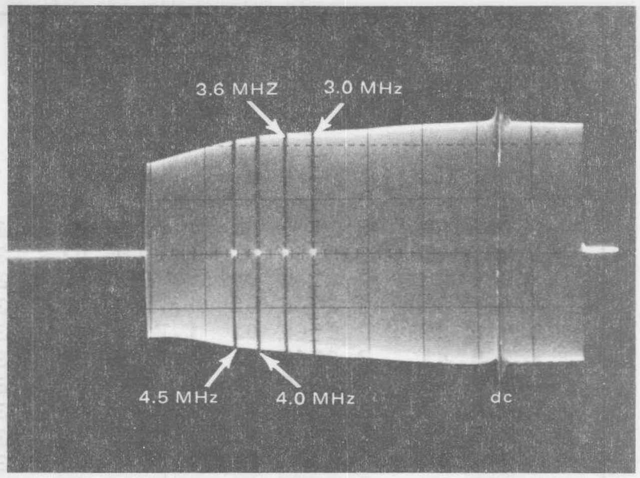
FIGURE 7g

	Curve 1a		Curve 1b		Curve 2a	
FREQUENCY (HZ)	V1 MAG(LIN)	V1 PHASE	V1 MAG(LIN)	V1 PHASE	V1 MAG(LIN)	V1 PHASE
1.00E 05	2.256E 01	1.798E 02	2.255E 01	1.787E 02	1.902E 01	1.797E 02
1.26E 05	2.256E 01	1.798E 02	2.255E 01	1.783E 02	1.902E 01	1.797E 02
1.58E 05	2.256E 01	1.798E 02	2.255E 01	1.779E 02	1.902E 01	1.796E 02
2.00E 05	2.256E 01	1.797E 02	2.254E 01	1.773E 02	1.902E 01	1.794E 02
2.51E 05	2.256E 01	1.796E 02	2.252E 01	1.766E 02	1.902E 01	1.793E 02
3.16E 05	2.256E 01	1.795E 02	2.250E 01	1.758E 02	1.902E 01	1.791E 02
3.98E 05	2.256E 01	1.794E 02	2.246E 01	1.747E 02	1.902E 01	1.789E 02
5.01E 05	2.256E 01	1.792E 02	2.241E 01	1.733E 02	1.902E 01	1.786E 02
6.31E 05	2.256E 01	1.790E 02	2.232E 01	1.716E 02	1.902E 01	1.783E 02
7.94E 05	2.256E 01	1.788E 02	2.218E 01	1.695E 02	1.901E 01	1.778E 02
1.00E 06	2.255E 01	1.784E 02	2.197E 01	1.668E 02	1.901E 01	1.772E 02
1.26E 06	2.255E 01	1.780E 02	2.164E 01	1.636E 02	1.900E 01	1.765E 02
1.58E 06	2.254E 01	1.775E 02	2.116E 01	1.596E 02	1.898E 01	1.756E 02
2.00E 06	2.253E 01	1.769E 02	2.045E 01	1.549E 02	1.896E 01	1.745E 02
2.51E 06	2.252E 01	1.761E 02	1.946E 01	1.495E 02	1.892E 01	1.731E 02
3.16E 06	2.249E 01	1.751E 02	1.815E 01	1.434E 02	1.887E 01	1.713E 02
3.98E 06	2.245E 01	1.738E 02	1.652E 01	1.369E 02	1.877E 01	1.691E 02
5.01E 06	2.239E 01	1.722E 02	1.465E 01	1.303E 02	1.863E 01	1.663E 02
6.31E 06	2.229E 01	1.702E 02	1.267E 01	1.239E 02	1.841E 01	1.628E 02
7.94E 06	2.214E 01	1.677E 02	1.070E 01	1.180E 02	1.808E 01	1.586E 02
1.00E 07	2.190E 01	1.647E 02	8.880E 00	1.128E 02	1.758E 01	1.535E 02

	Curve 2b		Curve 3a		Curve 3b	
FREQUENCY (HZ)	V1 MAG(LIN)	V1 PHASE	V1 MAG(LIN)	V1 PHASE	V1 MAG(LIN)	V1 PHASE
1.00E 05	1.902E 01	1.787E 02	1.394E 01	1.798E 02	1.393E 01	1.787E 02
1.26E 05	1.902E 01	1.784E 02	1.394E 01	1.797E 02	1.393E 01	1.784E 02
1.58E 05	1.901E 01	1.779E 02	1.393E 01	1.796E 02	1.393E 01	1.779E 02
2.00E 05	1.901E 01	1.774E 02	1.393E 01	1.795E 02	1.392E 01	1.774E 02
2.51E 05	1.899E 01	1.767E 02	1.393E 01	1.794E 02	1.391E 01	1.767E 02
3.16E 05	1.898E 01	1.759E 02	1.393E 01	1.793E 02	1.390E 01	1.759E 02
3.98E 05	1.895E 01	1.748E 02	1.393E 01	1.791E 02	1.388E 01	1.748E 02
5.01E 05	1.891E 01	1.735E 02	1.393E 01	1.788E 02	1.385E 01	1.735E 02
6.31E 05	1.884E 01	1.718E 02	1.393E 01	1.785E 02	1.380E 01	1.718E 02
7.94E 05	1.873E 01	1.698E 02	1.393E 01	1.781E 02	1.372E 01	1.697E 02
1.00E 06	1.856E 01	1.672E 02	1.393E 01	1.777E 02	1.359E 01	1.671E 02
1.26E 06	1.831E 01	1.640E 02	1.392E 01	1.771E 02	1.340E 01	1.639E 02
1.58E 06	1.792E 01	1.602E 02	1.392E 01	1.763E 02	1.312E 01	1.601E 02
2.00E 06	1.736E 01	1.556E 02	1.390E 01	1.753E 02	1.270E 01	1.555E 02
2.51E 06	1.657E 01	1.502E 02	1.389E 01	1.741E 02	1.212E 01	1.501E 02
3.16E 06	1.551E 01	1.442E 02	1.386E 01	1.726E 02	1.134E 01	1.440E 02
3.98E 06	1.418E 01	1.377E 02	1.382E 01	1.707E 02	1.036E 01	1.375E 02
5.01E 06	1.263E 01	1.309E 02	1.375E 01	1.684E 02	9.220E 00	1.308E 02
6.31E 06	1.097E 01	1.243E 02	1.364E 01	1.654E 02	7.998E 00	1.242E 02
7.94E 06	9.302E 00	1.182E 02	1.348E 01	1.618E 02	6.779E 00	1.181E 02
1.00E 07	7.738E 00	1.126E 02	1.323E 01	1.573E 02	5.637E 00	1.126E 02

FIGURE 7h

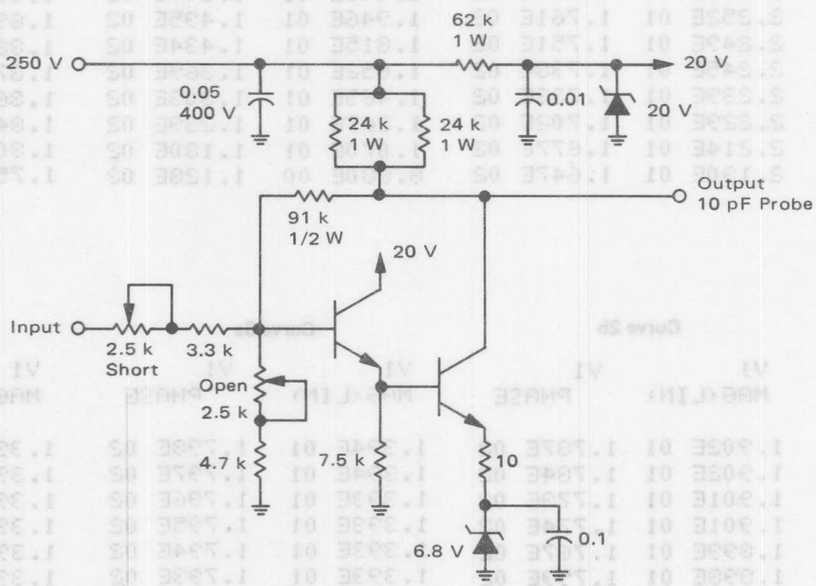




2 V/Div

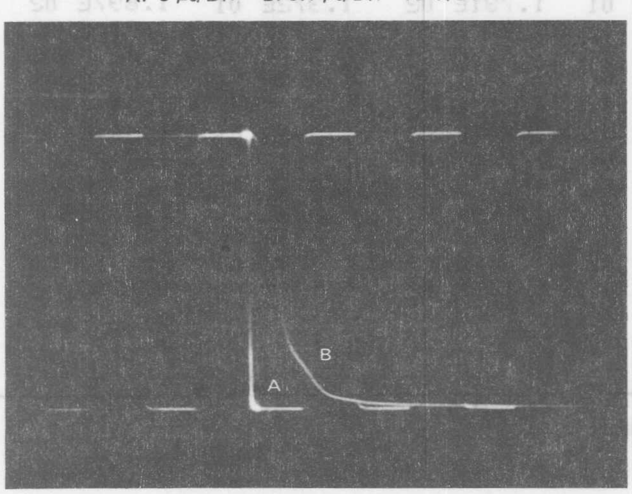
Sweep 0-6 MHz
Markers 3, 3.6, 4, 4.5 MHz

10 V/Div



100 kHz Squarewave Response
A. 5 μs/Div B. 0.1 μs/Div 10 V/Div

1 MHz Squarewave Response
Output 10 V/Div, Input 0.5 V/Div, 0.1 μs/Div



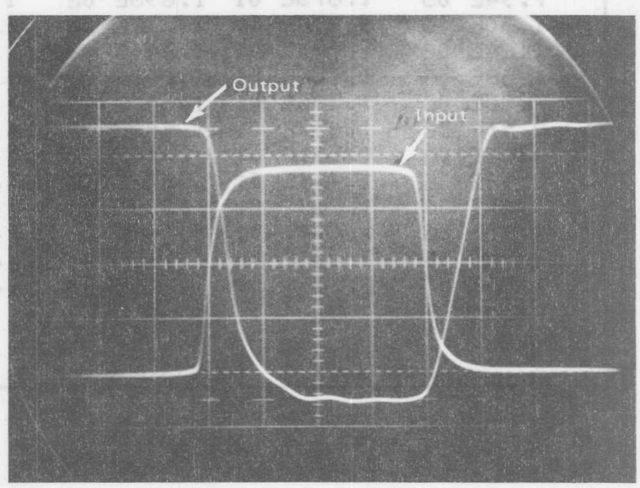
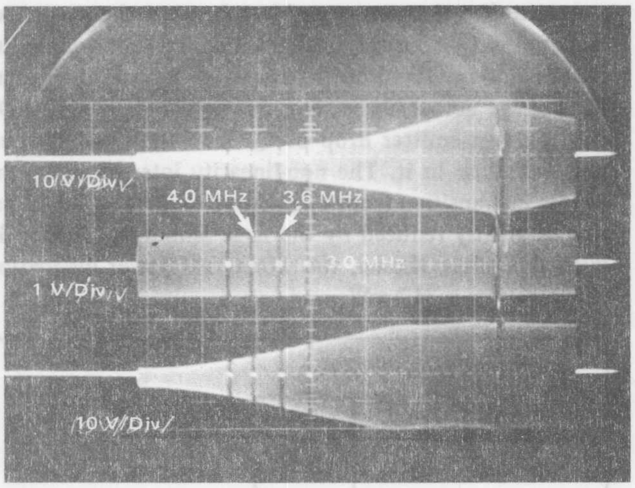


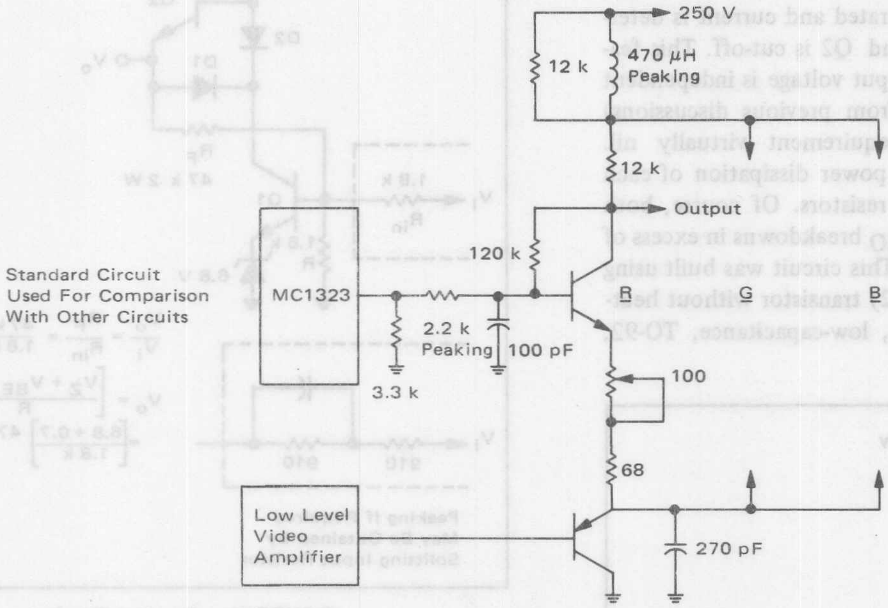
FIGURE 7i



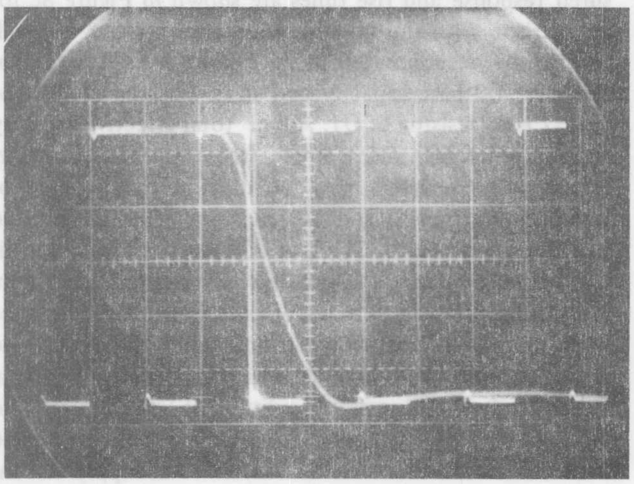
Output (No Peaking Components)
Uncompensated Response

Input (Injected Sweep @ PNP Base)

Output (With Peaking Components)
Fully Compensated Response



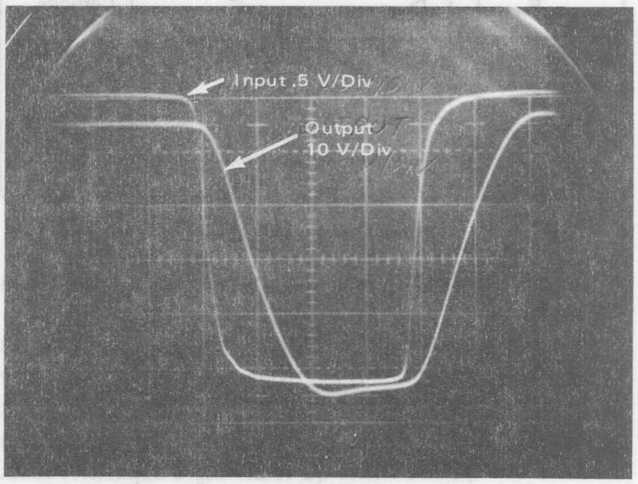
100 kHz Square Wave Response



10 V/Div

5 µs/Div and 0.1 µs/Div

1 MHz Square Wave Response



0.1 µs/Div

FIGURE 7j

ANOTHER NOVEL APPROACH

Motorola's European Division* has developed an output stage for the European market which has all the properties of our preceding discussion plus some added advantage and trade-offs. The basic configuration again uses a pseudo operational amplifier configuration as shown in Figure 8a. Here, an emitter-follower output is used (instead of the darlington), thus allowing a large load resistor on Q1, providing both high-voltage gain and low-output impedance. One very interesting property of this design is that it can be made to draw a constant current from the power supply if R_L and R_F are the same value. This becomes apparent if one assumes the input is approximately at ground and the output is at supply. Q2 (ignore base current) collector current is determined by R_F (V_{supply}/R_F) and a Q1 is cut-off. However, if the output is at ground, Q1 must be saturated and current is determined by R_L (V_{supply}/R_L) and Q2 is cut-off. This feature, plus the fact that the output voltage is independent of supply voltage (apparent from previous discussions) makes the power supply requirement virtually nil. Another advantage is the low power dissipation of each device due to the large load resistors. Of course, both devices must be capable of V_{CEO} breakdowns in excess of the maximum supply voltage. This circuit was built using the Motorola BF393 (MPS-A42) transistor without heatsinking. This is a high-voltage, low-capacitance, TO-92, 650-mW device.

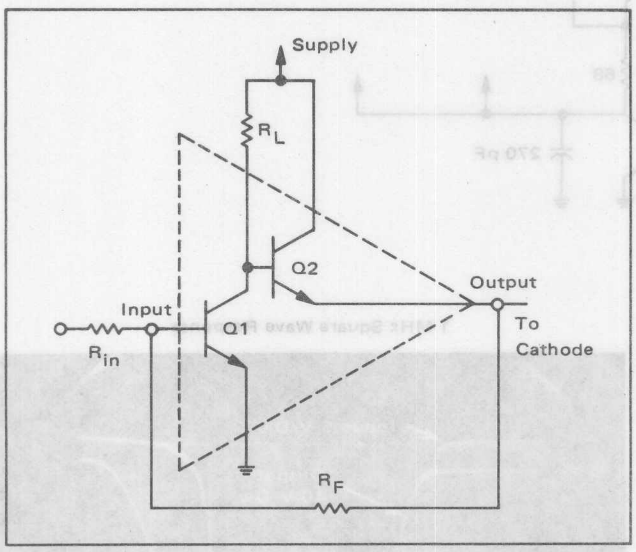


FIGURE 8a – Basic Circuit

A practical circuit using this configuration is shown in Figure 8b.

The equation for output voltage and gain are included in Figure 8b. The diode D1 provides a pulldown capability using Q1 as a current sink while driving capacitive loads at high slew rates. This is necessary since Q1 turning on tends to cut Q2 off when Q2 is driving a capacitive load. (Remember, output impedance of this configuration is low only while the amplifier is active.) Slewing in the opposite direction is no problem because the capacitive load tends to turn Q2 on harder as its base voltage increases during Q1 turnoff. During small-signal operation, the voltage across D1 is zero, due to the drop across D2 and the base-emitter drop of Q2, and therefore, no current will flow in it. The non-linearity introduced by the diodes in transition from small-signal to large-signal operation is greatly reduced by the large amount of feedback. With the values shown, the worst-case power dissipation per transistor is 312 mW.

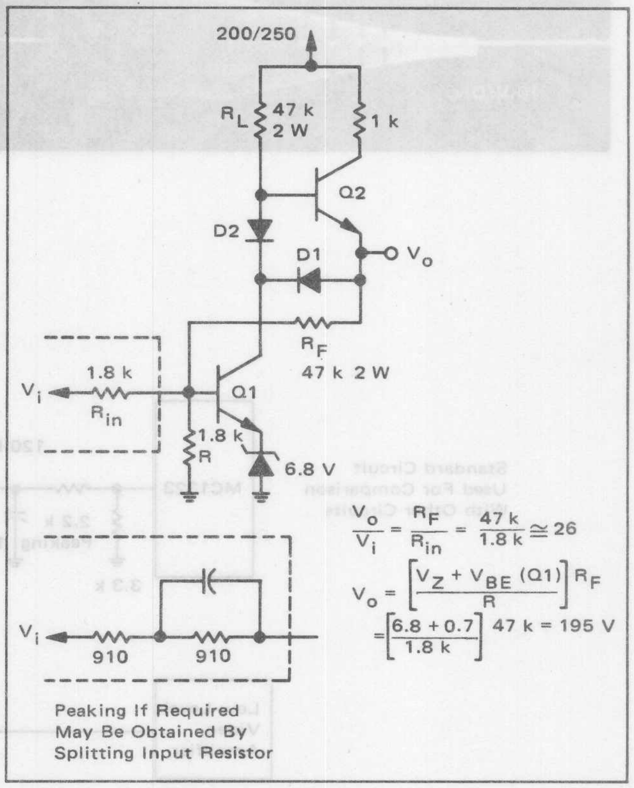


FIGURE 8b – Practical Circuit

The analysis of this stage will not be analyzed here because of its similarity to the preceding stage. The measured response and rise times are shown in Figure 8c for comparison with the other configurations. Note, the roll-off in the response which is caused by Q1 collector capacity, Q2 base capacity and stray working against the 47 kΩ load resistor of Q1. A 3 pF total effective capacity at that point will give a break frequency of 1.12 MHz. Lower value resistors may be used at the expense of transistor dissipation and power supply current required to extend bandwidth. The lack of voltage feedback at this internal point makes it susceptible to the transistor parameters and strays. This circuit offers excellent performance with minimal demand on supply and transistors and should prove to be a very advantageous circuit. Peaking in either the low-level video or splitting R_{in} (Figure 8b) and introducing a pole in the response at 1.12 MHz will yield 3 dB bandwidth in excess of 5 MHz. Total supply current per stage is 5.3 mA at a supply of 250 Volts. Variation of dc output voltage with a supply range of 200–250 Volts is 1.5 Volts, or less than 3%.

*Consumer Design Note CE-003 Europe "Innovative Video Output Amplifiers"

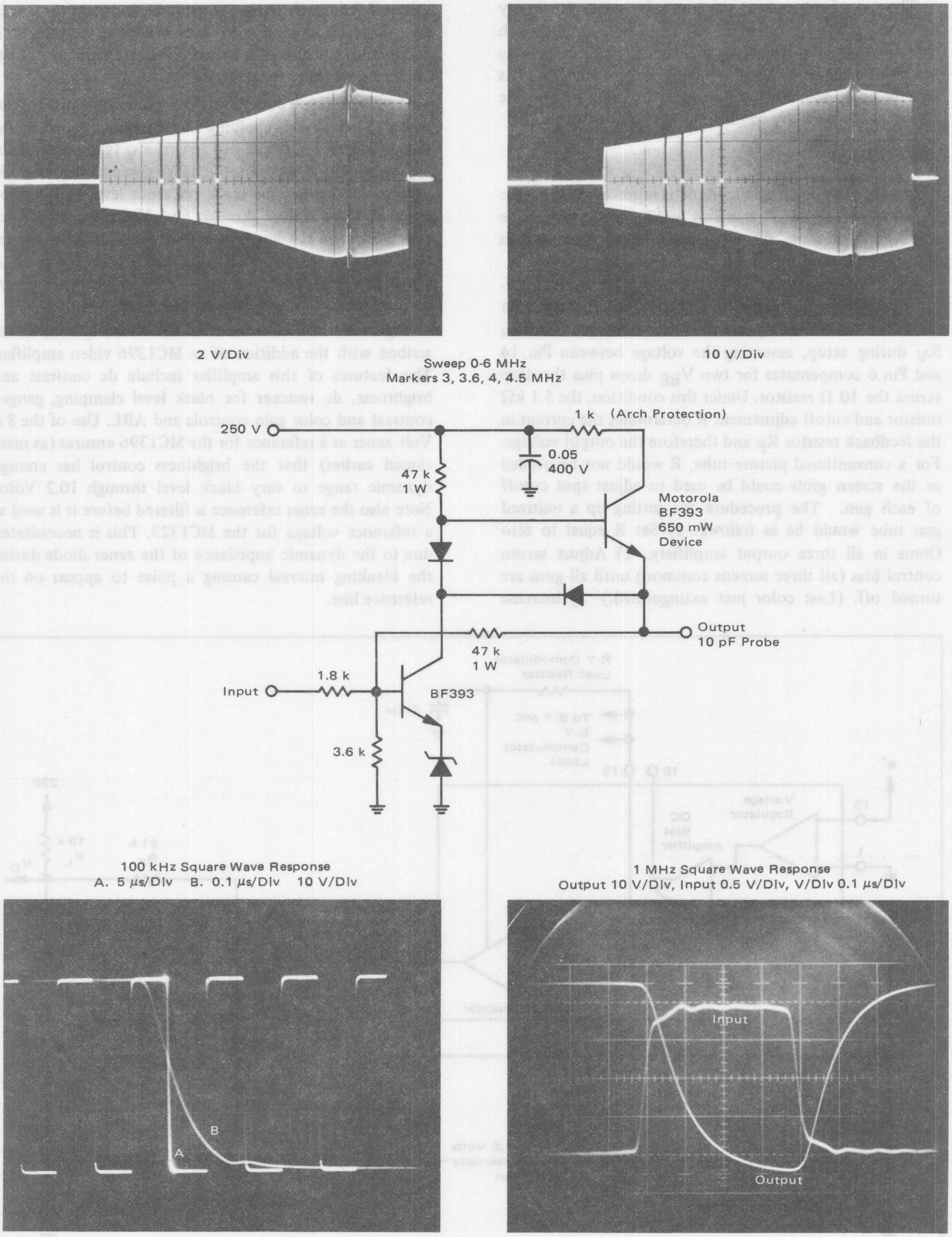

2 V/Div
10 V/Div
Sweep 0-6 MHz
Markers 3, 3.6, 4, 4.5 MHz
250 V
1 k (Arch Protection)
0.05
400 V
47 k
1 W
Motorola
BF393
650 mW
Device
Output
10 pF Probe
47 k
1 W
1.8 k
Input
BF393
3.6 k
100 kHz Square Wave Response
A. 5 μs/Div B. 0.1 μs/Div 10 V/Div
B
A
1 MHz Square Wave Response
Output 10 V/Div, Input 0.5 V/Div, V/Div 0.1 μs/Div
Input
Output

FIGURE 8c

The use of this output stage and the stages previously discussed will be covered in the next section. Although the previous configuration will be shown, everything described will apply to this output stage as well. As has already been perceived, this stage is also able to sum the chrominance signal by adding another input resistor.

THE SYSTEM

The circuit shown in Figure 6f is reproduced in Figure 9a for convenience. This circuit will be used as a vehicle for describing some system considerations because it is already familiar to the reader.

The luminance signal voltage source V1 is opened during black level adjustment, or picture tube gun cutoff, via service switch S1. This guarantees that no current flows in R_V during setup, assuming the voltage between Pin 14 and Pin 6 compensates for two V_{BE} drops plus the drop across the 10 Ω resistor. Under this condition, the 5.1 kΩ resistor and cutoff adjustment R determines the current in the feedback resistor R_F and therefore the output voltage. For a conventional picture tube, R would not be needed as the screen grids could be used to adjust spot cutoff of each gun. The procedure for setting up a unitized gun tube would be as follows: 1) Set R equal to zero Ohms in all three output amplifiers. 2) Adjust screen control bias (all three screens common) until all guns are turned off. (Last color just extinguished.) 3) Increase value of R in each output stage independently until each gun turns on. As value of R is increased, current I1 is decreasing, causing a corresponding decrease in voltage across R_F and thereby reducing V_O.

As a final step in completing the setup procedure, service switch S1 is closed, applying the luminance signal to the output stage. To ensure not having R_V (gain control for adjusting color temperature) change gun cutoffs, it is only necessary to ensure the black reference level of the video signal (V1) be shifted through 10.2 Volts, via the brightness control in the low-level video amplifier. This guarantees that no current will flow in R_V at black level. A low-impedance drive source for V1 also eliminates drive control interaction and chroma matrixing.

Figure 9b is a schematic of the entire system as described with the addition of an MC1396 video amplifier. The features of this amplifier include dc contrast and brightness, dc restorer for black level clamping, ganged contrast and color gain controls and ABL. Use of the 8.8 Volt zener as a reference for the MC1396 ensures (as mentioned earlier) that the brightness control has enough dynamic range to vary black level through 10.2 Volts. Note also the zener reference is filtered before it is used as a reference voltage for the MC1323. This is necessitated due to the dynamic impedance of the zener diode during the blanking interval causing a pulse to appear on the reference line.

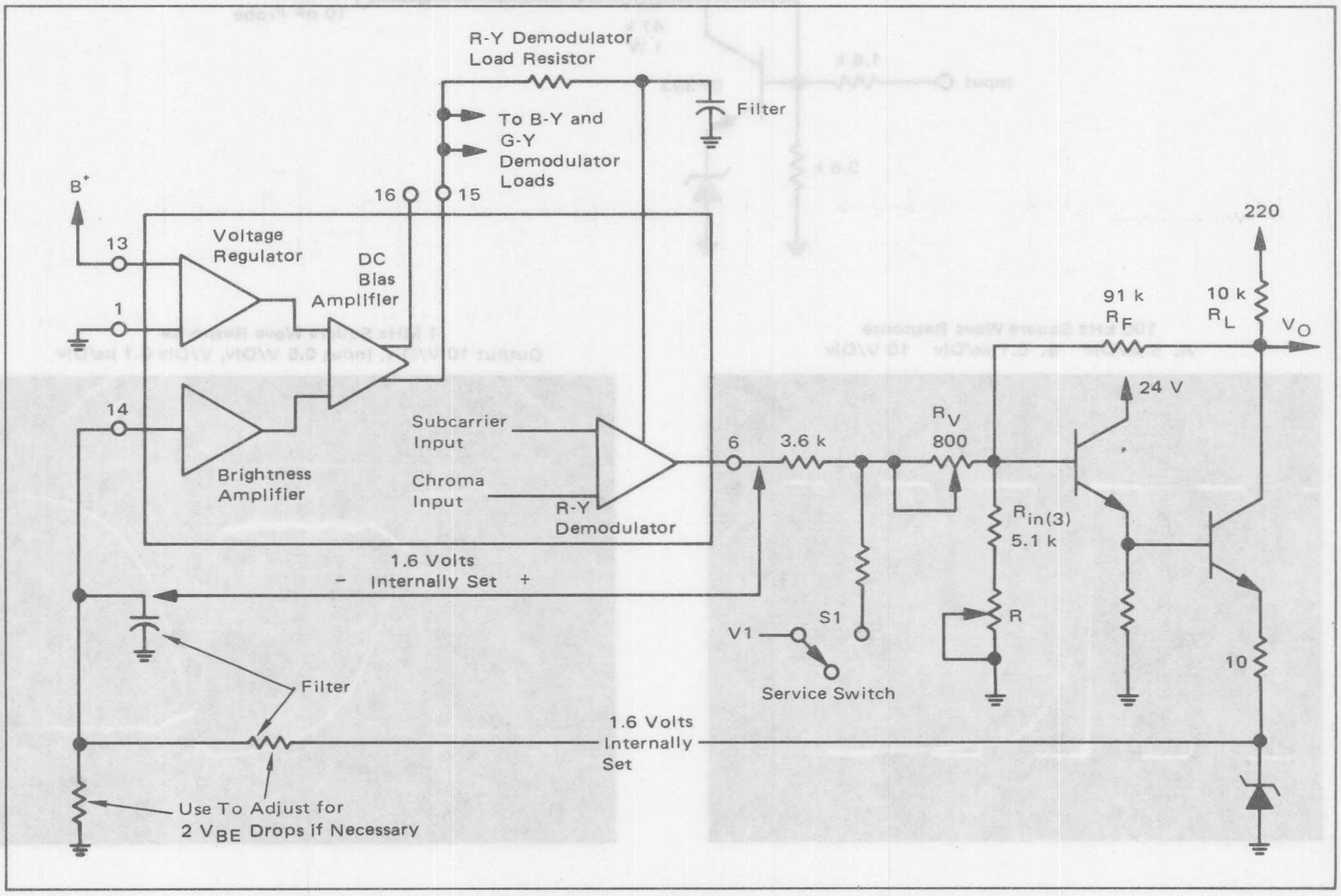


FIGURE 9a

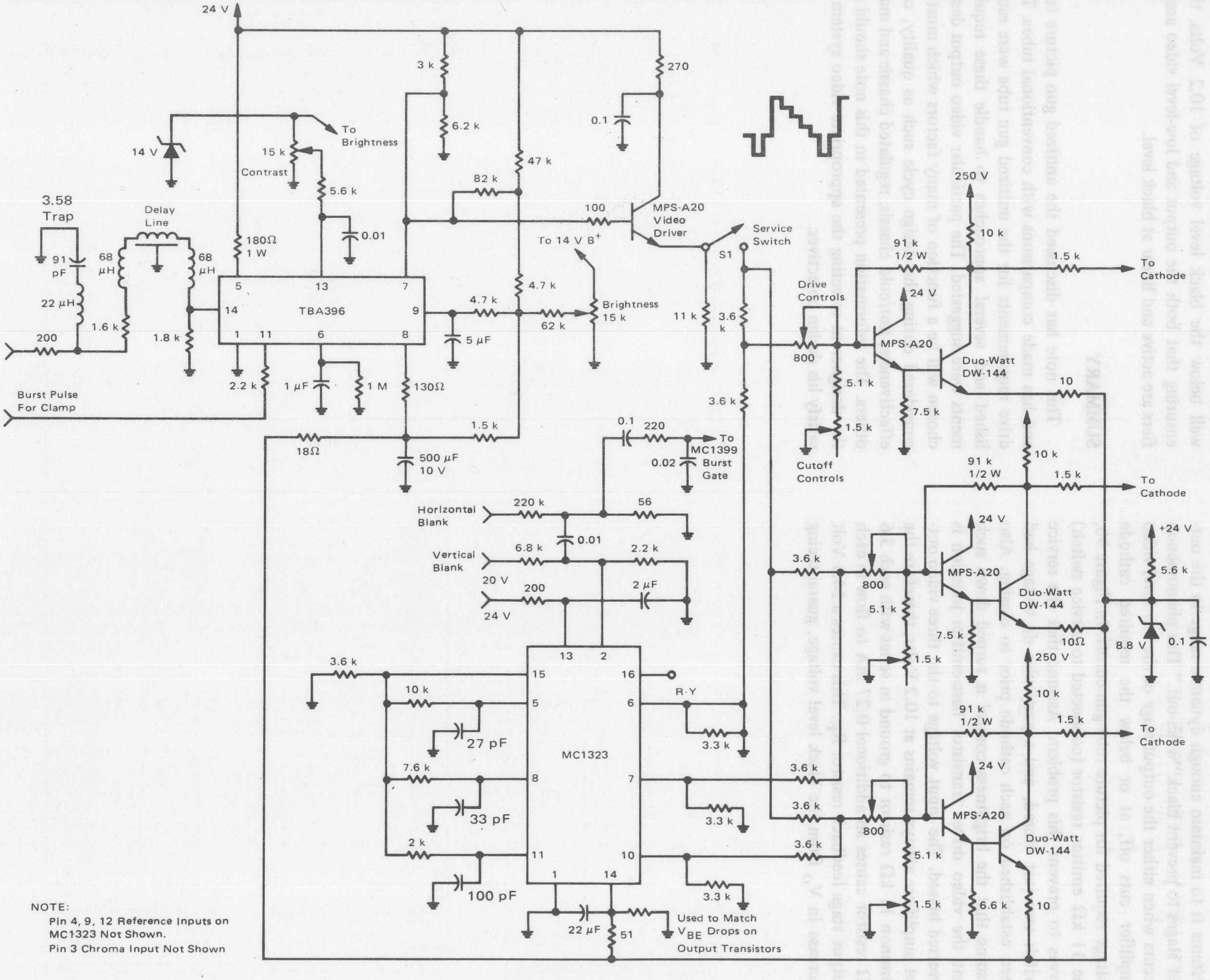


FIGURE 9b – Video System

It is beyond the scope of this paper to go into any detail on the TBA396. More information may be obtained by contacting the Consumer Applications Department.

Another important consideration in the design of video systems is to maintain enough dynamic range in the output stages to prevent black "wash-out." This phenomenon occurs when either the output stage or the low-level video amplifier cuts off, at or below the required cathode voltage required for picture tube gun cutoff. In Figure 9b, the 11 kΩ emitter resistor (connected to service switch) serves to prevent this problem. Assume that the service switch was just closed, and a particular dc voltage had been established on each cathode prior to closing. Also assume that the brightness control is turned down such that the video driver transistor base-emitter junction is reversed biased. The input voltage to the three video output amplifiers always remains at 10.2 Volts, therefore the common 11 kΩ resistor to ground in series with each 3.6 kΩ resistor causes an additional 0.27 mA to flow in each output stage feedback resistor R_F. This causes a 24.5 Volt increase in V_o from the black level voltage, guaranteeing cutoff. The resulting voltage on top of the 11 kΩ resistor due to the 0.27 mA from each output stage is 8.9 Volts. As the brightness control is now turned up, the drive transistor starts to conduct at 8.9 Volts at the emitter, well below the black level voltage of 10.2 Volts, thus ensuring that both the output and low-level video amplifiers are active and linear at black level.

SUMMARY

This note has discussed the unitized gun picture tube and has made comparisons with conventional tubes. The drive requirements for the unitized gun tube were established and several approaches to handle these requirements were suggested. The particular video output design chosen will be a function of many factors which must be considered during the design cycle such as quality, cost effectiveness, hot/cold chassis, regulated chassis and many others. The information presented in this note should aid the designer in selecting the appropriate video system to satisfy his design objectives.

APPENDIX

SMALL-SIGNAL CURRENT GAIN TRANSFER CHARACTERISTICS

TRANSISTOR RESPONSE

PART LIST

** RESISTORS **

NAME	FROM	TO	MINIMUM	NOMINAL	MAXIMUM
RL	2	3	1.000E 00	1.000E 00	1.000E 00

** INDEPENDENT CURRENT SOURCES **

NAME	FROM	TO		MINIMUM	NOMINAL	MAXIMUM
IIN	0	1	DC VALUE	2.500E-05	2.500E-05	2.500E-05
			AC MAGNITUDE	1.000E-05	1.000E-05	1.000E-05
			PHASE	.000E 00		

** INDEPENDENT VOLTAGE SOURCES **

NAME	FROM	TO		MINIMUM	NOMINAL	MAXIMUM
VS	0	2	DC VALUE	1.900E 02	1.900E 02	1.900E 02

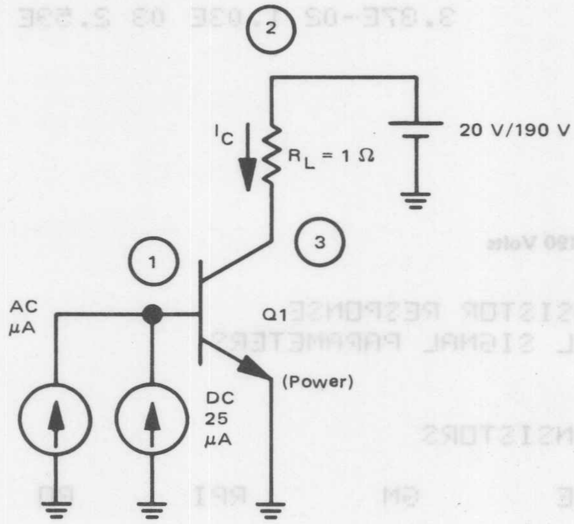
** BIPOLAR JUNCTION TRANSISTORS **

NAME	C	B	E	DEVICE MODEL	AREA FACTOR
Q1	3	1	0	PWR	1.000

DEVICE NAME = PWR

TYPE = NPN

PARAMETER	VALUE
BF	4.000E 01
BR	1.000E 00
RB	.000E 00
RC	.000E 00
RE	.000E 00
CCS	2.000E-12
TF	.000E 00
TR	.000E 00
CJE	3.000E-11
CJC	1.000E-11
IS	1.000E-14
PE	1.000E 00
PC	1.000E 00
VA	.000E 00
EG	1.110E 00



V_S = 20 Volts

```
TRANSISTOR RESPONSE
SMALL SIGNAL BIAS SOLUTION              TEMPERATURE     27.000 DEG C

 ITERATIONS  =   6

 NODE  VOLTAGE      NODE  VOLTAGE      NODE  VOLTAGE      NODE  VOLTAGE

(  1)        .6548 (  2)    20.0000  (  3)     19.9990

 VOLTAGE SOURCE CURRENTS

 NAME        CURRENT

 VS          1.000E-03 AMPS

 TOTAL POWER DISSIPATION   2.00E-02  WATTS

 TRANSISTOR OPERATING POINTS

 NAME      MODEL        IB        IC        VBE      VBC      VCE     BETADC

 Q1        PWR       2.51E-05  1.00E-03    .655  -19.344   19.999    40.0


TRANSISTOR RESPONSE
SMALL SIGNAL PARAMETERS                 TEMPERATURE     27.000 DEG C

 TRANSISTORS

 NAME       GM        RPI       RO        CPI       CMU      BETAAC   FT

 Q1      3.87E-02 1.03E 03 2.59E 12 3.98E-11 2.22E-12    40.0 1.46E 08
```

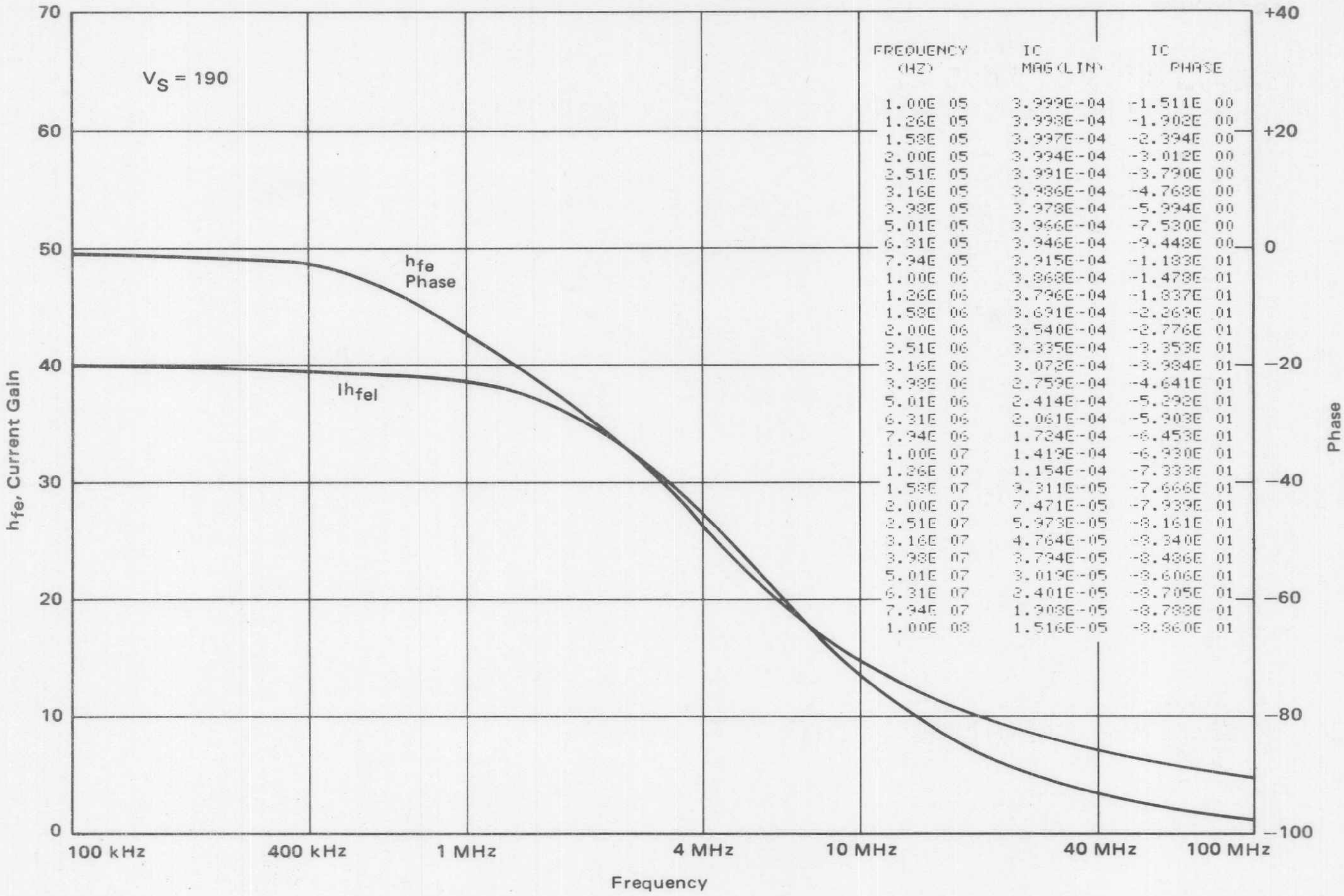
V_S = 190 Volts

```
TRANSISTOR RESPONSE
SMALL SIGNAL PARAMETERS                 TEMPERATURE     27.000 DEG C

 TRANSISTORS

 NAME       GM        RPI       RO        CPI       CMU      BETAAC   FT

 Q1      3.87E-02 1.03E 03 2.59E 12 3.98E-11 7.25E-13    40.0 1.52E 08
```



FREQUENCY (HZ)	IC MAG(LIN)	IC PHASE
1.00E 05	3.999E-04	-1.511E 00
1.26E 05	3.998E-04	-1.902E 00
1.58E 05	3.997E-04	-2.394E 00
2.00E 05	3.994E-04	-3.012E 00
2.51E 05	3.991E-04	-3.790E 00
3.16E 05	3.986E-04	-4.768E 00
3.98E 05	3.978E-04	-5.994E 00
5.01E 05	3.966E-04	-7.530E 00
6.31E 05	3.946E-04	-9.448E 00
7.94E 05	3.915E-04	-1.183E 01
1.00E 06	3.868E-04	-1.478E 01
1.26E 06	3.796E-04	-1.837E 01
1.58E 06	3.691E-04	-2.269E 01
2.00E 06	3.540E-04	-2.776E 01
2.51E 06	3.335E-04	-3.353E 01
3.16E 06	3.072E-04	-3.984E 01
3.98E 06	2.759E-04	-4.641E 01
5.01E 06	2.414E-04	-5.292E 01
6.31E 06	2.061E-04	-5.903E 01
7.94E 06	1.724E-04	-6.453E 01
1.00E 07	1.419E-04	-6.930E 01
1.26E 07	1.154E-04	-7.333E 01
1.58E 07	9.311E-05	-7.666E 01
2.00E 07	7.471E-05	-7.939E 01
2.51E 07	5.973E-05	-8.161E 01
3.16E 07	4.764E-05	-8.340E 01
3.98E 07	3.794E-05	-8.486E 01
5.01E 07	3.019E-05	-8.606E 01
6.31E 07	2.401E-05	-8.705E 01
7.94E 07	1.908E-05	-8.788E 01
1.00E 08	1.516E-05	-8.860E 01